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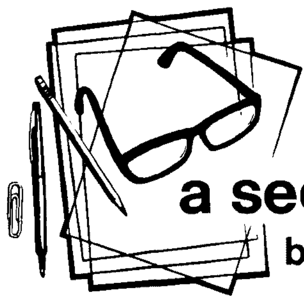


a second look

by Jim Fisk

Although this new, larger edition of *ham radio* may seem like a nuisance if your shelves are designed for the old size, I think you'll soon discover that the advantages of the larger format far outweigh the slight inconvenience of storage problems. For one thing, the larger page size allows us to present larger schematics, so there will be less segmented drawings than there have been in the past. If you're building a project or tracing out a circuit diagram, switching from one page to another as you go through a schematic can be annoying, and often leads to wiring errors. The larger page size also means that the photographs will be larger, so you will be able to more clearly see how the author layed out his original circuit.

However the graphical advantages of the larger page size are small potatoes when compared to the big bonus of the larger format: more available space for technical and construction articles. This 104-page issue, for example, contains as much reading material as any two of our previous issues – the more than 50 pages of technical articles in this issue, if scaled down to the old size, would fill nearly 90 pages. This not only means that we've got to work twice as hard to keep *ham radio* filled with the kind of articles you want to read, it also means that we can provide more basic construction articles and tutorial material that we didn't have room for in the old format. While we will continue to publish the latest technical developments in amateur radio, the increase in editorial space will allow us to broaden our horizons to include features which will appeal to a wider audience. Some of those new features are included in this issue – others will be added in the months ahead.

One of those new features is *repair bench*, a monthly column devoted to troubleshooting your own equipment. We have had many requests for such a column but until now, because of the nearly weekly changes in modern communications technology, there simply wasn't room in the magazine. The first few *repair bench* columns will be geared to basic troubleshooting techniques, while future columns will attack such subjects as receivers, transmitters, ssb equipment, vhf fm, RTTY, logic systems, slow-scan television and much more. The column won't be written by one person, but by different authors who have proven expertise on the topic covered by a particular column. Although I have several authors already lined up, I'm looking for others with troubleshooting experience who would be interested in writing some columns. If you have suggestions for topics, or would like to contribute, please drop me a line.

The *circuits and techniques* column which we have published irregularly for the past several years will once again become a monthly feature beginning with this issue. *Circuits and techniques* will also take on a different look in the coming months as we use it as a vehicle for presenting new circuits, technical developments and construction techniques which come to our attention. If you develop a simple circuit for a special task, are using a well-known circuit in an unusual application, or run across an interesting circuit or technique in a foreign publication, we'd like to hear about it.

The popular *ham notebook* column which we've been publishing since 1968 will continue to be a monthly feature, as will the *microprocessor* column which we introduced last month. We're also looking for amateur-oriented construction projects which are built around microprocessor chips.

We have several other new features being developed which will be published in the coming months. One of these will be the *weekender*, a simple project that can be built in a few hours time over one weekend. A unique feature of the *weekender* is that we will arrange to have all components and a circuit board available from one easy-to-reach commercial source. The first of the *weekender* projects is scheduled for publication in the February issue, and we're busily rounding up future *weekender* candidates from our authors. If you have a project which you think might qualify as a *weekender*, we will be glad to consider it for publication. Suggestions for future *weekender* projects are also welcome – we may be able to place your idea in the hands of an author who can come up with a finished product.

Our editorial staff is very excited about the many possibilities of the new, larger size, and we're looking forward to making *ham radio* bigger and better than ever before. Your comments and suggestions are always welcome.

Jim Fisk, W1DTY
editor-in-chief

publisher's log

skip tenney
W1NLB

A look through this issue will quickly show that *ham radio magazine* is at a significant turning point in its eight year history. This is by far the biggest magazine we've ever published. Not only are the pages larger, but it also has far more editorial matter, more columns, more color and yes, even more ads than ever before.

This change is typical of what's happening throughout Amateur Radio. The whole hobby appears to be at a turning point which will lead to many changes over the next few years which could well make today's Amateur world seem quite unfamiliar.

When we started only eight years ago VHF fm was unknown to most Amateurs. Slow-scan TV was in its very infancy. Almost no Amateur gear was solid-state at the time, while digital concepts and integrated circuits were virtually unheard of in amateur work.

Now we suddenly find ourselves at a new starting point as digital techniques are coming at us in a rush led by the exciting new microprocessor chips which are scheduled to change much of our daily world as they take charge of your kitchen, automobile and workplace. It goes without saying that their effect over the next few years on even a relatively simple Amateur station will be significant.

Arriving almost simultaneously with the birth of *ham radio magazine* were the long awaited rules outlining *Incentive Licensing*, which have provided the basic framework of the Amateur licensing structure for the past seven years.

Again during the past year the Amateur community has had an excellent opportunity to debate at length another major step in our regulatory history commonly known as *Restructuring*. At this writing it appears that within the next few months these new ideas will become reality, but possibly in a very different form than originally proposed just a year ago, but definitely including the much discussed no-code license. The concept of *Reregulation* has also been introduced by Commission officials and should further influence regulations by which we must conduct ourselves.

The Amateur Radio business community has also seen many changes. Your all time high acceptance of Amateur products is permitting many exciting and useful new products to be offered which would have been out of the question just a few years ago. Attention to our advertising pages in the months ahead will show many outstanding surprises waiting for you.

Both *ham radio magazine* and *hr report* will be right there in the middle of these many exciting new developments and will bring them to you step by step as they unfold. We'll be doing our best to show you what is happening and just what can be done to insure that both you and Amateur Radio realize maximum benefit from these many changes.

Skip Tenney, W1NLB
publisher

presstop

JOHN JOHNSTON REPLACED Prose Walker as chairman of the World Administrative Radio Conference Amateur Radio working group at the group's December meeting in Washington. Prose had been a prime mover in getting this very important activity organized and going, and despite his retirement from the FCC in July had headed up its September meeting during the ARRL national convention in Reston and seemed likely to continue with it on a consultant basis with the FCC. However, the staffing and budget crunch in the Amateur and Citizens Division brought on by the CB landslide workload put a crimp into those plans so Division Chief John Johnston will be taking it over.

Though Prose' Expertise will certainly be missed, John is expected to carry the WARC preparation effort ahead with minimal interruption. Under Prose' direction the basic organization had been firmly established and the various task force chairmen and their groups moving along nicely, so the transition should be a relatively painless one.

900 MHZ AMATEUR BAND is receiving consideration both in and out of the FCC. The recent opening of 115 MHz of spectrum in that region to commercial two-way users will accelerate technical development in that frequency range, and Amateur Radio (and/or possibly Class E CB) has at least a chance to pick up a portion of the remainder.

Amateur Space And Satellite Communications would find a new band in the 900 MHz region particularly valuable — it's high enough to get away from a lot of noise and low enough that atmospheric absorption is not a problem. The possibility has already been explored in WARC meetings and a proposal for such a band will probably become a WARC group recommendation.

OSCAR 7 is being seriously affected by users putting signals much stronger than needed into it on Mode B. Overloading is causing excessive battery discharge and may be responsible for mode switching and shutdowns. Area coordinators and others are asked to watch for signals causing "pumping," report calls of offenders to W3HUC c/o AMSAT and advise those nearby of their abuses.

Demonstrating Mode B Sensitivity, W6CG made over 20 contacts in one week running 500 mW to a dipole antenna! Bud's QSOs included Hawaii and Maryland.

REPEATER CROSSBANDING, DOCKET 20113, has been approved and became effective December 15. Report and order will permit unlimited crossband operation of repeaters in the authorized repeater subbands, covers several related topics.

Definition Of "Automatic Retransmission" has been added to the rules, characterizes an "automatic retransmission" as one initiated by a received signal. Automatic retransmission is restricted to repeaters, auxiliary links or remotely controlled stations such as a remote base which has an auxiliary link station as a part of its system. In the latter case, the remote base is limited to retransmitting the signals of its auxiliary link station only.

PAPERWORK FOR REPEATERS and other remotely-controlled Amateur stations will be simplified greatly by an FCC action adopted in November. As of December 1, technical showings will no longer be a required part of the license application for such a station and technical information will be required only as part of the permanent station log. Repeater license applicants, for example, need only specify that their proposed station will be remotely controlled. System block diagrams, control details and the like need not accompany the application but must be entered in the permanent station log. Similarly, repeater-control stations will not even need to specify what repeater they intend to become a control station for — that's entered in their control station log and the log of the repeater they control.

Net Result of this important change is to speed up license processing greatly since technical evaluation will no longer be a part of the license granting procedure.

Note That All Information previously required as part of the FCC file record is still required in the permanent log. This is spelled out in new Part 97.103b, which replaces 97.41c and 97.41e. Control station requirements are spelled out in new Part 97.103d.

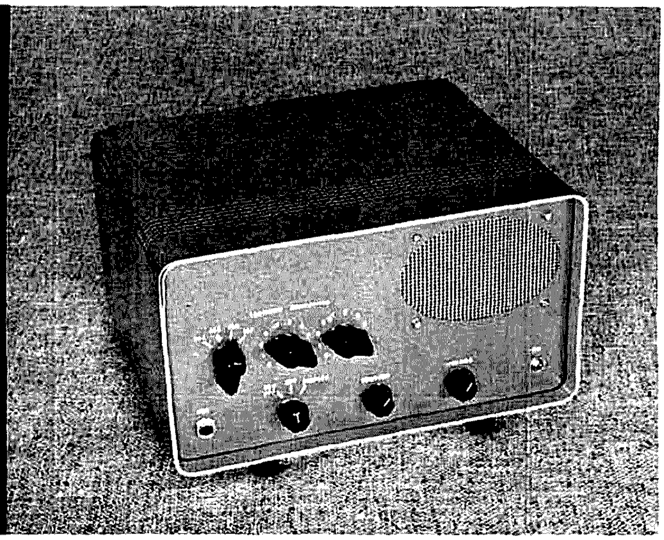
Prohibition Against Portable Or Mobile operation of a remotely controlled station in Part 97.88e has also been deleted. However, during portable or mobile operation a positive control system is still required and the usual requirements for ID, logging and notification must still be observed. Note too that the prohibition against portable or mobile operation of auxiliary link stations has not been relaxed.

REPEATER AND CLUB STATION TRUSTEES should be aware that group organization plans and constitutions are being checked by FCC legal people to be sure funding of group Amateur stations is not in violation of Part 97.112. All new applications are checked as a matter of course, and files on old licenses are sometimes pulled for review on a random basis. Groups whose fund raising systems seem to ask money for operation or use of the station are very likely to be cautioned.

BICENTENNIAL PREFIX LIST in last May's Presstop had a typo which should be corrected. WN1-WNØ can use AK1-AKØ — not AG1-AGØ as shown. Use of the alternate bicentennial prefixes is entirely optional, but remember that they don't go into effect until 0500Z January 1, 1976, and are good until January 1, 1977.

ALL IRCS IN CIRCULATION will be honored for first class overseas postage regardless of date of issue through the end of 1976, according to latest post office info. After that all earlier IRCs will be void and only latest issue will be valid.

CANADA GOES AFTER IGNITION NOISE with a new Radio Interference Regulation that takes effect next September 1. The new regulation will severely limit the permissible radiation from any spark ignition engine, includes autos, chain saws and snow mobiles, with the one exception of aircraft engines. The regulation will eventually be extended to include other RFI sources such as power tools and high voltage transmission lines.



synthesized two-meter fm transceiver

Frequency heterodyne techniques, synthesizer modulation and modular construction are combined in this novel design

This article describes a two-meter fm transceiver containing a 400-channel frequency synthesizer. The transceiver is designed to operate from a 12-volt dc source. By using heterodyne techniques rather than frequency multiplication, only one frequency at a time is generated by the synthesizer, which is used for both transmit and receive modes. Lock-up problems are avoided by eliminating the need to generate the offsets in the synthesizer. With the heterodyne scheme, the synthesizer changes frequency directly in 10-kHz increments, which greatly simplifies its design; you need only dial in the desired transmit frequency along with the desired receive mode. The receiver offset, ±600 kHz for repeater or zero kHz for simplex operation, is generated by a separate crystal-controlled oscillator. An interesting feature of the transceiver is that the modulation is applied directly to the synthesizer, which results in excellent-quality audio with simple circuitry.

general description

Fig. 1 shows the functional elements. The synthesizer tunes from 12.01 to 16 MHz in 10-kHz steps. Modulation is applied directly to the voltage-controlled oscillator (VCO) control line from a clipper preamp. In the receive mode, the clipper preamp is disabled by switching the B+ line. The VCO output is buffered after which the signal is split and fed to two double-balanced mixers,

By Robert W. Wilmarch, W1CMR, and William R. Wade, K1IJZ, Roberts Road, Wellesley, Massachusetts 02181

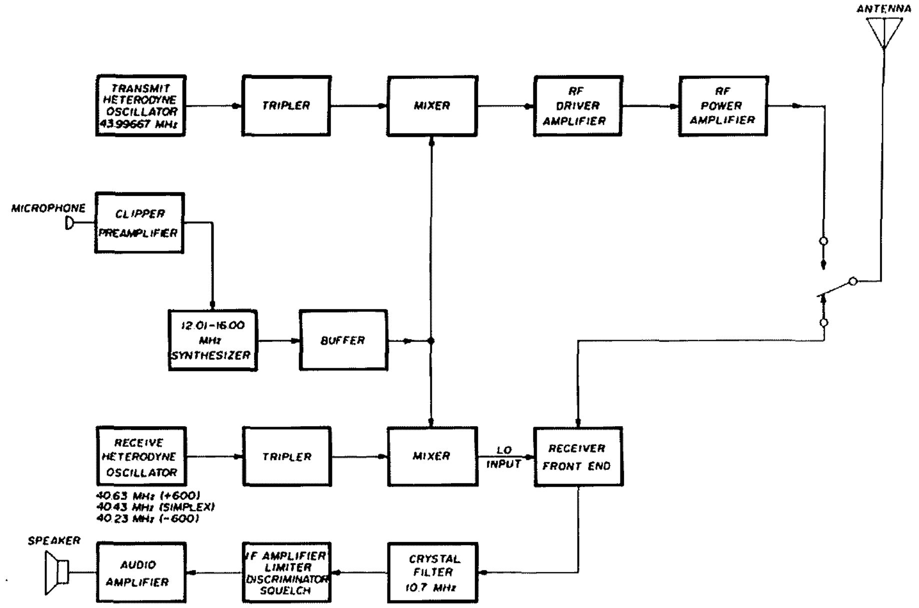


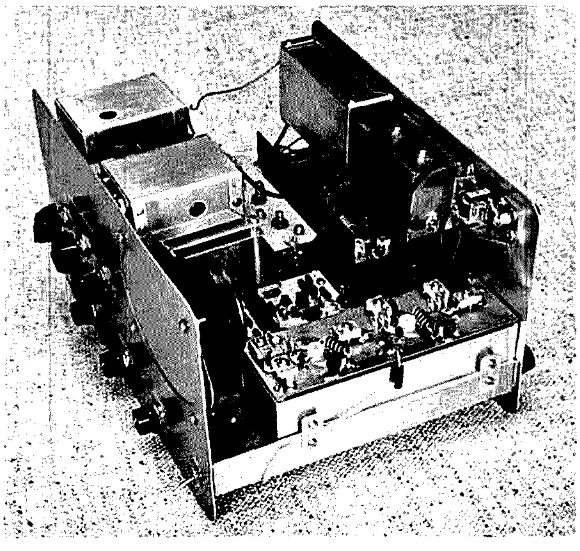
fig. 1. Block diagram of the synthesized 2-meter fm transceiver.

one in the transmit line; the other in the receive line. In both cases the synthesizer signals are fed into the local oscillator (LO) ports of the mixers.

In the transmit mode a signal at 131.99 MHz is added to the synthesizer signal so that the resultant signal covers 144 to 148 MHz. The front-panel controls are marked to indicate the transmit frequency. The mixer output is fed into a driver amplifier where the signal level is raised to about 2 watts. This signal in turn drives a 15 to 20 watt power amplifier. In the receive mode the transmit heterodyne oscillator and mixer stage are disabled through the B+ line. Voltage is left on the driver and power amplifier stages since these stages are run in class C and consume negligible power without drive.

On receive, the synthesizer signal is mixed with one of three crystal-controlled frequencies depending on the desired operating mode. The resulting sum frequency is the LO frequency required to heterodyne the receive frequency to the (nominal) 10.7 MHz intermediate frequency. This i-f signal is fed through a crystal filter which determines receiver selectivity. The circuits following the filter are conventional. In the transmit mode only the audio amplifier is disabled, again through the B+ line. Two small relays are used. One switches the antenna from receive to transmit. The other, in the B+ line, turns various circuits on and off as described above. A double-pole, double-throw relay may be used for switching.

Overall view. The audio amplifier and optional power amplifier are shown immediately behind the speaker.



synthesizer

The synthesizer generates frequencies between 12.01 and 16.00 MHz in 10-kHz steps. During transmit this output is heterodyned with a 131.99-MHz signal to produce transmit frequencies between 144.00 and 147.99 MHz. On receive, the required LO frequency is obtained by heterodyning the synthesizer output with either 121.29 MHz for simplex operation, 121.89 MHz for normal repeater operation, or 120.69 MHz for reverse repeater operation. Because of the heterodyning scheme, this synthesizer is simplicity itself. It requires none of the 1-count detectors, out-of-lock detectors, or count offset circuits of synthesizers used in *multiplier* service.[1]

Above 7 MHz the programmable divider chain of SN74192s swallows a count due to propagation delays. This action causes a 1 count (10 kHz) offset in the synthesizer output frequency from that to which the divider is set. This offset is compensated in the heterodyne process to yield the correct transmit or LO frequency with respect to dial setting at the mixer output.

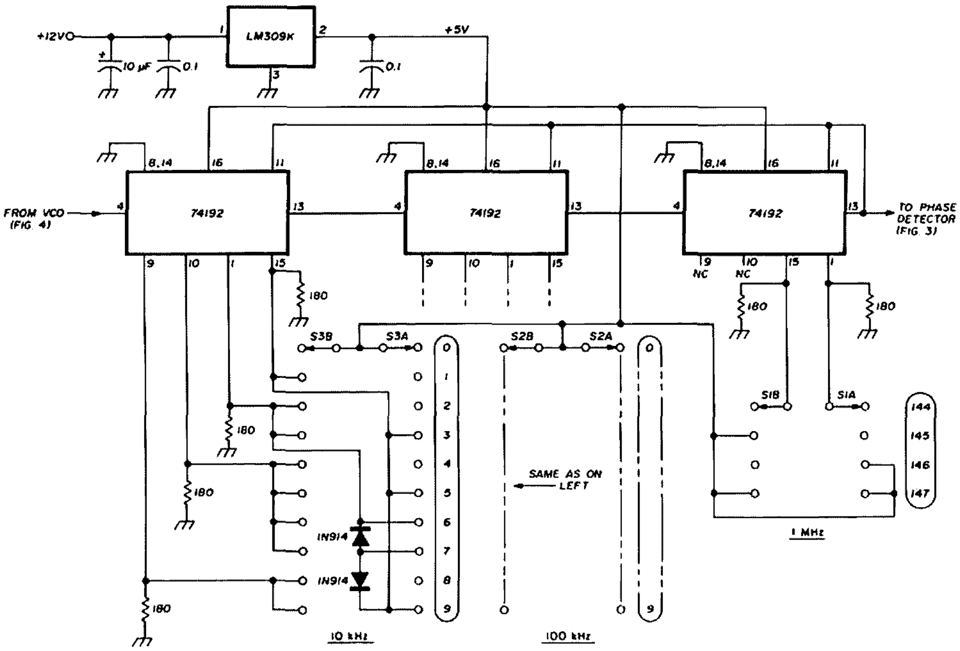
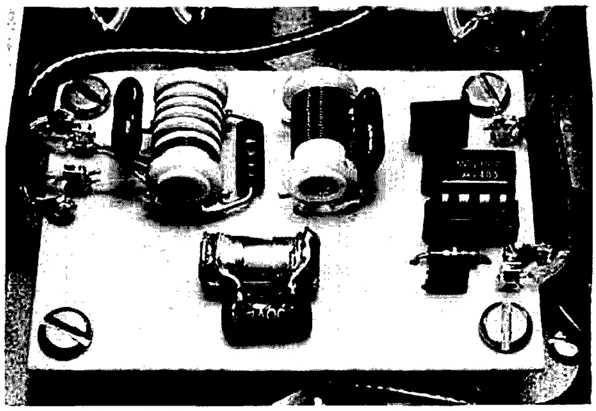


fig. 2. Programmable divider.

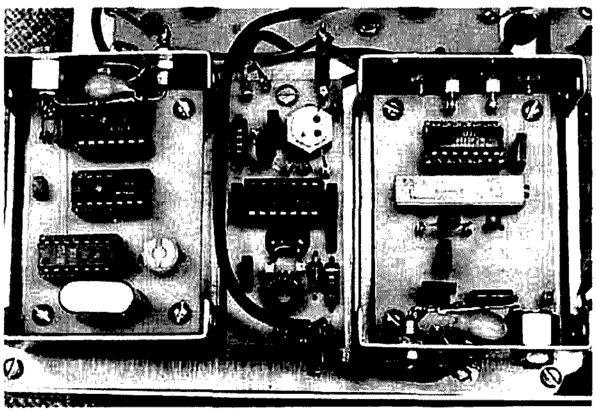
The programmable divider, fig. 2, is unique in that the ÷12 through ÷15 functions are obtained from a single decade counter chip. This bit of magic is accomplished by using the last 74192 as a downcounter, which is preset to a 12, 13, 14, or 15. The decade limitation on the 74192 holds only in an upcount mode. The other two counters are presettable from 0 to 9, and the string of three 74192s divides the VCO frequency by a number between 1201 and 1600 with the programmed inputs set between 1200 and 1599.

The phase detector and filter, fig. 3, are straight from the MC4044 data sheet with an extra capacitor on the filter output to help supress the 10-kHz ripple on the VCO control line. Adjustment of the 10k pot in the filter is accomplished by listening to the VCO on a receiver, tuning 10 kHz off to find a VCO sideband and tweaking the pot for minimum sideband signal.



Synthesizer buffer amplifier.

Top of synthesizer showing, left to right, phase detector and filter; VCO, and reference oscillator.



The VCO, fig. 4, is an LC oscillator using the MC1648 as the active element. This circuit proved superior in performance to any of the available multivibrator type VCOs. Watch out for the MV1401! It's an expensive ($9) wide-range varicap, and again it proved superior to the less-expensive diodes. A glance at the synthesizer schematics shows that the phase detector, reference oscillator, programmable divider, and VCO each has its own LM309 5-volt regulator. A regulator is mandatory in phase-locked loops to decouple the circuits from each other. Any modification of this decoupling scheme should be avoided. Usual RC and LC decoupling techniques do not compare with the use of three-terminal regulator ICs.

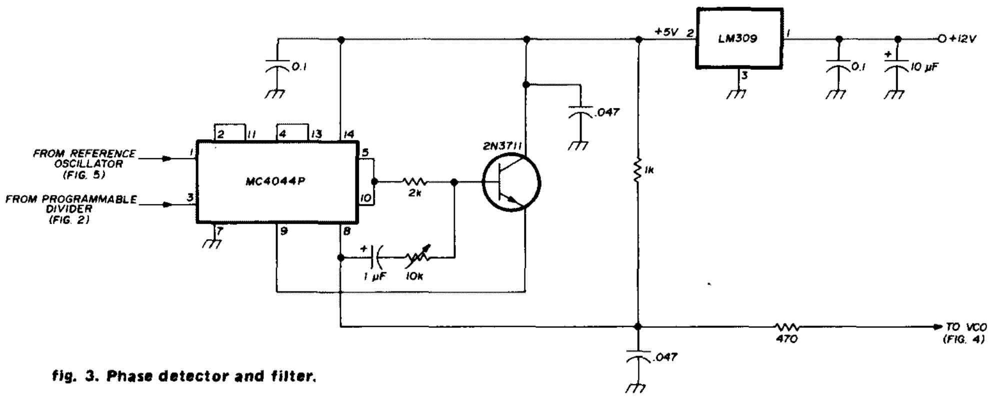


fig. 3. Phase detector and filter.

The programmable divider is constructed on double-sided board with the wiring side at ground and the component side at +5 volts. The V_{cc} pins of the IC sockets are bent out and soldered directly to the 5-volt side, while the ground leads are brought through the boards and soldered directly to the foil. This approach provides a low impedance V_{cc} line, which prevents possible erratic synthesizer behavior.[2]

The reference oscillator, fig. 5, is a 1-MHz crystal oscillator followed by two decade dividers to yield the 10-kHz reference frequency. The synthesizer output is buffered as shown in fig. 6. A double-tuned output circuit provides a flat response over the full 4-MHz range. A single-tuned output stage will suffice if the transceiver is set up to operate over a 2-MHz range. In this case the MHz switch may be replaced with a single-pole, single-throw switch.

From the VCO buffer amplifier the signal is split and fed to two separate mixer stages. These stages, (fig. 7), are identical except for minor differences in the tuned circuits. In each case, the stage is used to add the synthesizer output to that of a heterodyne oscillator. In one case the sum frequency is the transmit frequency, while in the other it is the receive LO frequency. Double-balanced mixers are used because they happened to be available. Suitable mixers may be built[3] or purchased for about $7.00 new and perhaps for considerably less on the surplus market. A mixer stage using a dual gate 40673 mosfet was tried with apparently satisfactory results; however, the suppression of the other mixing product was not verified. Other approaches should work equally well.[4]

Care was taken to provide 50-ohm terminators to each mixer port. The synthesizer buffer is fed into the LO port, and with the coupling arrangement shown, the buffer provides an LO signal of +7 dBm. The heterodyne oscillator signals were adjusted by varying the position of the output links so that the power at the mixer was near zero dBm. These adjustments did not appear to be critical. By using an in-line layout for the mixers, no instabilities were encountered.

The receive mixer stage is powered at all times, while the transmit mixer stage and its heterodyne oscillator are powered only during transmit, which is necessary to prevent a receiver birdie in the simplex mode.

modulator

The first attempts at modulating the transmitter were along conventional lines; the modulating voltage was applied to a tuning diode in the transmit heterodyne oscillator crystal circuit. While this method worked, the audio quality left something to be desired. After a number of attempts to improve matters, this approach was abandoned in favor of directly modulating the VCO in the synthesizer. The results were indeed gratifying, with reports of excellent audio quality. Full deviation is

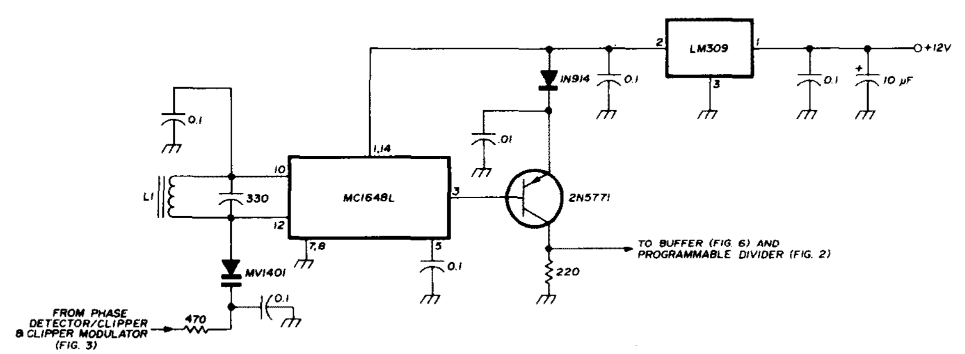


fig. 4. Voltage-controlled oscillator. L1 is 4 turns on Amidon T50-6 toroidal core.

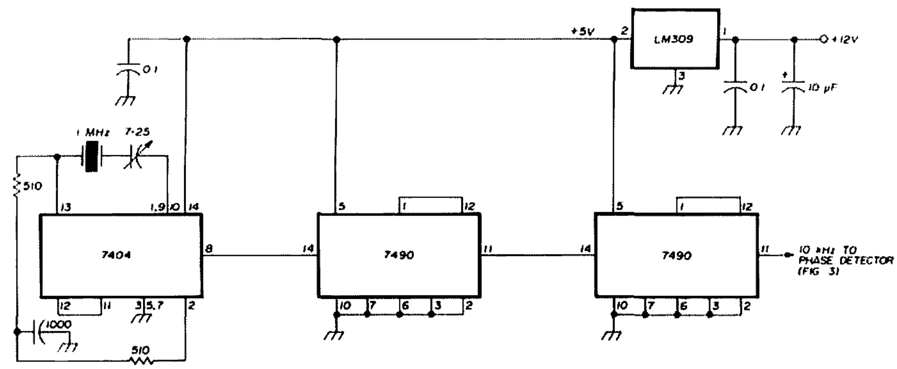
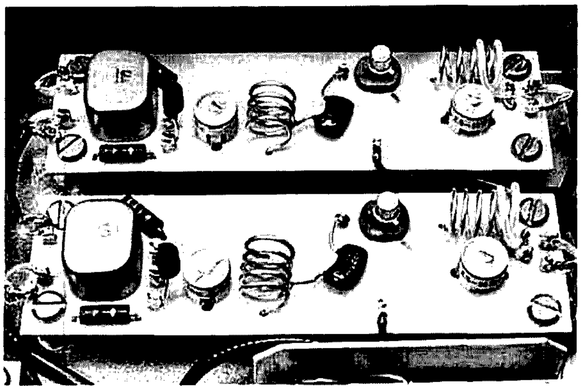


fig. 5. Reference oscillator.

obtained with only a few millivolts of modulation superimposed on the VCO line. This signal level is many times below that required to disturb the phase-locked-loop stability.

The clipper preamp, (fig. 8) is a modification of a circuit originally designed to modulate a tuning diode[5] where several volts of modulation were required. Since only millivolts are now required, the output stage was changed to a simple emitter follower, eliminating several components.



Transmit and receive mixers.

heterodyne oscillators

The heterodyne oscillators are shown in figs. 9 and 10. The circuits differ only in the number of crystals and the addition of a zener regulator in the transmit oscillator. Overtone crystals in the 40-MHz range are used. The second stages are conventional triplers using a mosfet to minimize oscillator loading. Tripler stage output is through a one-turn link.

During tuneup, remember that a final frequency is the sum of the synthesizer frequency and that of the heterodyne oscillator. The reference oscillator should be adjusted first, then the transmit heterodyne oscillator, to produce the desired output transmit frequency. A frequency counter is recommended for this procedure. The receive heterodyne oscillator crystals should be adjusted by tweaking their series capacitors for best received audio.

receiver front end

The receiver front end (fig. 11) is similar to a circuit described in 1968.[6] Only minor changes were made in the rf and mixer stages. The original fets were replaced with 40673s, and the mixer output matches a crystal filter. Gate-protected fets eliminate the need for diodes at the antenna. With gate protection no special precautions are necessary in handling these transistors; however, the 3N128 is not protected and care must be exercised. The mixer output impedance is determined primarily by the resistor across the output tank and is chosen to match the crystal filter.

The front end and i-f stages show a direct connection to the crystal filter. This is fine if the physical layout is close and there is no dc return in the filter. If a dc return is present, a blocking capacitor must be used to prevent

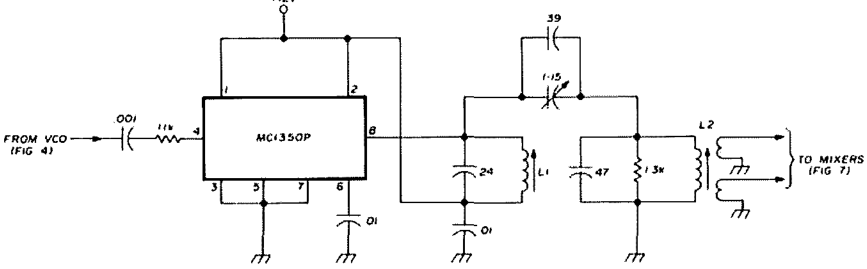
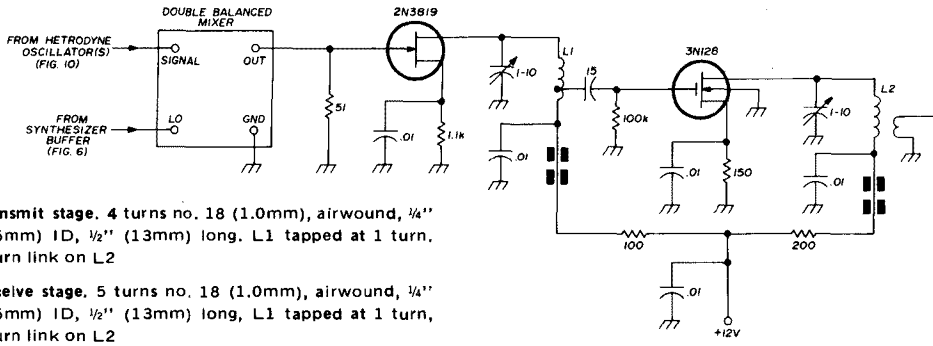


fig. 6. Synthesizer buffer amplifier. L1 and L2 are 20 turns no. 28 (0.3mm) on ¼" (6.5mm) form. Output links on L2 are each 5 turns.



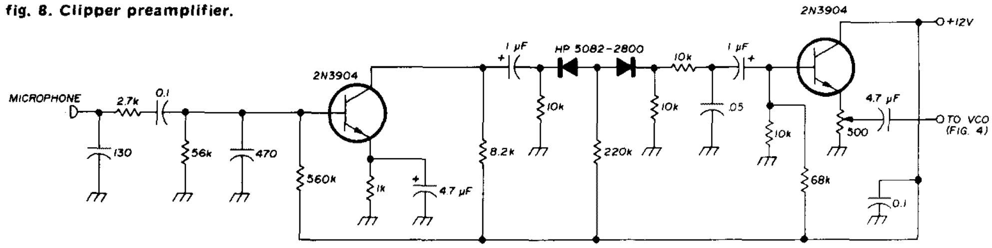
L1,L2 **Transmit stage.** 4 turns no. 18 (1.0mm), airwound, ¼" (6.5mm) ID, ½" (13mm) long. L1 tapped at 1 turn, 2-turn link on L2

L1,L2 **Receive stage.** 5 turns no. 18 (1.0mm), airwound, ¼" (6.5mm) ID, ½" (13mm) long, L1 tapped at 1 turn, 2-turn link on L2

fig. 7. Mixer. One each is required for transmit and receive. L1 and L2 are 4 turns each on the transmit board and 5 turns each on the receive board.

short circuiting the mixer drain voltage. If the circuits are to be separated physically, a coaxial line must be used for shielding. In this case the line becomes part of the mixer tank circuit, thus requiring a smaller tank capacitor than the 56 pF shown. A good rule of thumb for coaxial cable is 30 pF per foot. The LO input from the receive mixer stage is amplified and applied to the 40673 front-end mixer through a 5-pF coupling capacitor. Signal level should be about 1.5 volts.

fig. 8. Clipper preamplifier.

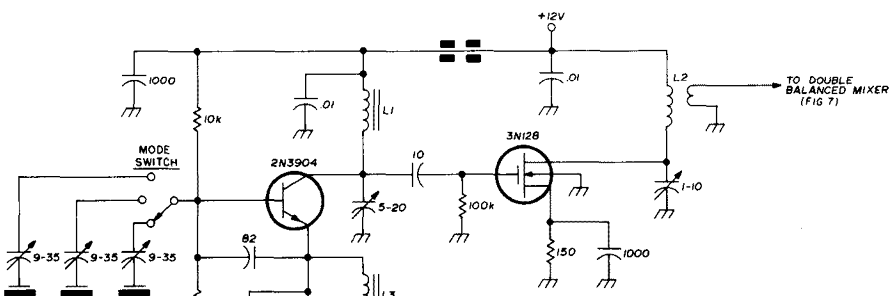


i-f subsystem and audio

The receiver is a single-conversion device. While single conversion has certain advantages, the trick is to recover sufficient audio in narrowband fm service. The RCA CA3089E linear integrated circuit is a complete fm i-f subsystem (**fig. 12**). While this IC was designed for wideband use, it's possible to realize 290 millivolts of recovered audio for +5-kHz deviation,[7] which is ample to drive the audio stage to full output. The squelch control operates smoothly with this device and doesn't have the annoying pop-out characteristics as in some circuits. A tuning or signal-strength meter may be used with the circuit. However, it was decided not to include this feature. Instead, pin 13 is used as a test point for receiver front-end tuneup.

The Q of the quadrature coil across pins 9 and 10

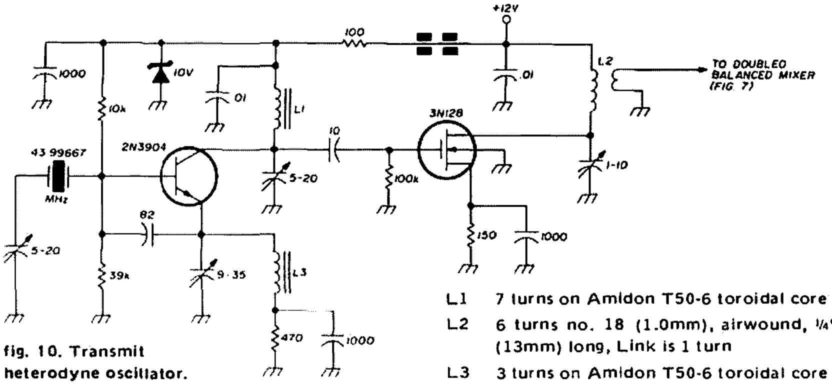
fig. 9. Receive heterodyne oscillator.



L1 7 turns on Amidon T50-6 toroidal core

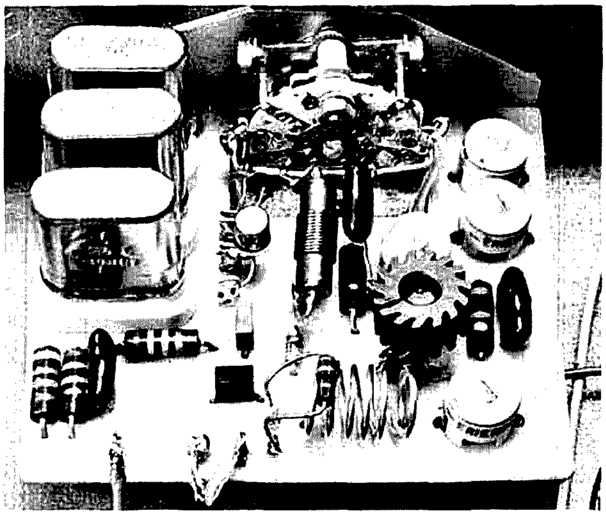
L2 6 turns no. 18 (1.0mm), airwound, ¼" (6.5mm) ID, ½" (13mm) long. Link is 2 turns

L3 3 turns on Amidon T50-6 toroidal core



L1	7 turns on Amidon T50-6 toroidal core
L2	6 turns no. 18 (1.0mm), airwound, 1/4" (6.5mm), 1/2" (13mm) long, Link is 1 turn
L3	3 turns on Amidon T50-6 toroidal core

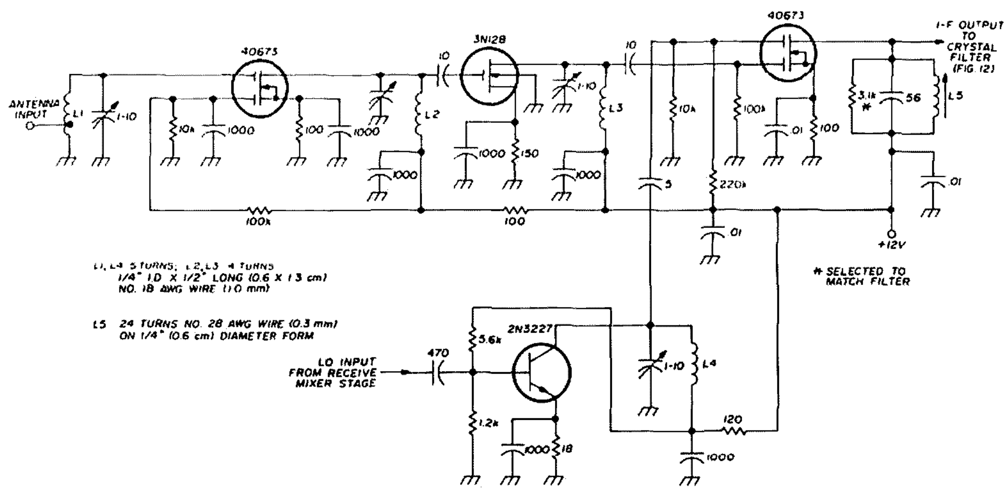
fig. 10. Transmit heterodyne oscillator.



Receive heterodyne oscillator.

determines, to a great degree, the level of recovered audio. The RCA test results[7] were for a Q of 120. Using an available core, a value of 220 was measured, which resulted in somewhat more recovered audio. This value of Q, however, was reduced to a value consistent with the sensitivity of the audio amplifier by simply padding the coil with a suitable resistor. The effect is to greatly reduce the sensitivity of the circuit to temperature changes. Stability of this stage may be checked by looking at pin 9 with an oscilloscope. If the circuit is oscillating, a square wave will be seen.

A 2N3819 junction fet matches the crystal filter to the CA3089E. Because of the very high input impedance of this transistor, the filter load resistor from gate to ground is chosen according to the requirements of the filter. Stage gain isn't critical and need not be more than necessary to overcome the filter insertion loss. Any audio stage with sufficient sensitivity may be used. The MFC9020 is a 2-watt amplifier requiring only 200 millivolts of drive.



L1,L4	5 turns no. 18 (1.0mm), airwound, 1/4" (6.5mm) ID, 1/2" (13mm) long	L2,L3	4 turns no. 18 (1.0mm), airwound, 1/4" (6.5mm) ID, 1/2" (13mm) long
		L5	24 turns no. 28 (0.3mm), on 1/4" (6.5mm) slug-tuned form

fig. 11. Receiver front end.

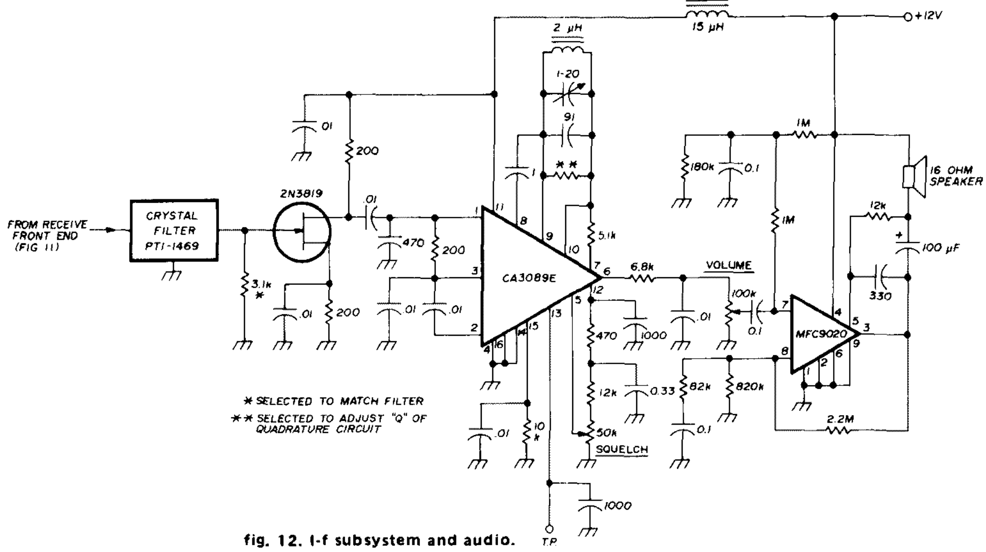


fig. 12. I-f subsystem and audio.

The rf driver amplifier, **fig. 13**, is conventional and is driven directly from the transmit mixer stage. Output is about 2 watts and the circuit will match a 50-ohm load. If 2 watts is sufficient, the output chain may be terminated at this point. For additional power, an amplifier such as the *VHF Engineering* unit shown in the photo provides output in the 15 to 20-watt range.

construction

Standard copper-clad board and point-to-point wiring are used. A minimum of tools are required and the difficulty of making printed boards is avoided. An advantage of the modular approach is that a circuit can be completed, tested, and set aside until the overall unit is ready for assembly. All interconnecting lines use small coaxial cable where length is not critical, which permits flexibility of the final layout. Rotary wafer switches were chosen for the frequency select controls. While significant space saving can be achieved by using BCD-coded thumbwheel switches, a rotary format affords a definite ease of operation.

This transceiver has given trouble-free operation for about a year with excellent signal reports. While heterodyning, digital-frequency synthesizers, and synthesizer modulation are all well-known processes, the combination of these features offers an attractive approach to those who like to try something different.

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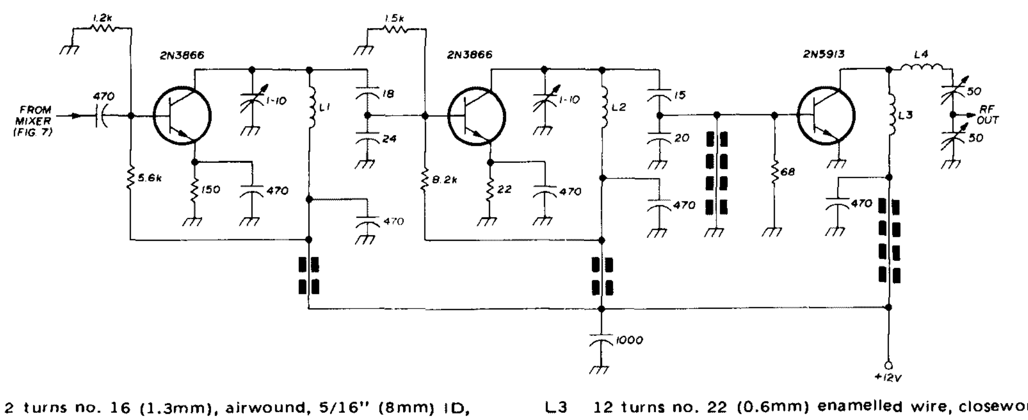
ham radio



L1,L2 2 turns no. 16 (1.3mm), airwound, 5/16" (8mm) ID, 3/16" (5mm) long

L3 12 turns no. 22 (0.6mm) enamelled wire, closewound on 3/16" (5mm) mandrel

L4 5 turns no. 16 (1.3mm), airwound, 5/16" (8mm) ID, ½" (13mm) long

fig. 13. Rf driver amplifier.

six digit 50-MHz frequency counter

A frequency counter has several advantages over a frequency standard. Instead of listening and tuning for crystal-oscillator harmonics on a receiver, a counter can provide a direct readout in frequency from the signal being measured. An instrument such as this can be a very valuable asset for the amateur who likes to build his own variable-frequency oscillators, transmitters, and receivers. With this frequency counter I was able to align a homemade crystal filter for an ssb rig, using the counter to pinpoint the exact location of the filter passband. When the counter is used with signal generators, precision alignment of amateur equipment is a snap.

The frequency counter described here and shown schematically in fig. 1 is designed for use in the hf spectrum to 50 MHz, with a signal at the input having an amplitude of about 50 mV rms. The digital readout displays the frequency in kHz with resolution to the nearest 100 Hz. Construction cost of the counter is about $50 including the power supply and cabinet. The cost will be lower if the ICs are in your goodie bin. Printed circuit boards are not available for this project. The entire counter, with the exception of the power supply, was built on perfboard – the kind with holes on a 0.1-inch (2.5mm) grid.

The goal of this article is to present a working design for a high-performance instrument that requires a minimum of ICs. However, I'd like to offer some observations based on experience with the project. I've noticed that the 50-MHz response is largely device-dependent. I had to select SN74S00 devices to get the counter to squeak up to 50 MHz. The ICs used in the counter were obtained from Poly Paks, as was the SN74196 decade counter. With the prescaler circuits published in *ham radio*[1,2,3] the counter should work well into the 432-MHz range.

The heart of the counter is a crystal-controlled oscillator. This 1-MHz source is a free-running multivibrator made up of two NAND gates (U1A and U1B) with a crystal as the frequency determining element. The 220-ohm resistors bias the gates in a class-A amplifier condition so that the oscillator is self starting and sustaining. The remaining two gates in the quad NAND package are used as buffers to isolate the oscillator from the loading effects of the IC stages that follow. U1D provides a buffered 1-MHz output to a BNC jack on the rear panel of the counter. The 1-MHz output is a very close approximation of a square wave, rich in harmonics, and provides a means of checking the oscillator with WWV. It also can be used for checking out the counter itself. If the 1-MHz output is coupled to the input jack, the counter will display 1000.0 kHz. The trimmer capacitor in series with the crystal is used in the zero beating process.

The frequency counter performs by sampling the input signal for a finite period of time. For example, if we were to couple a 1-MHz signal to a chain of decade counter stages for exactly one second, then 1 million pulses will have been counted. If the sampling time is reduced two orders of magnitude to 0.01 second, then the counter will register 10,000 pulses. Thus if 10,000 pulses are counted for each 1 MHz, the least-significant digit on the counter would represent 100 Hz. It's easy to see the importance of having a device that will perform the function of gating the unknown frequency with great precision.

The time-base divider chain is composed of four cascaded decide counters (U2-U5) followed by a flip-flop

By Jim Pollock, WB2DFA, 6 Terrace Avenue, New Egypt, New Jersey 08533

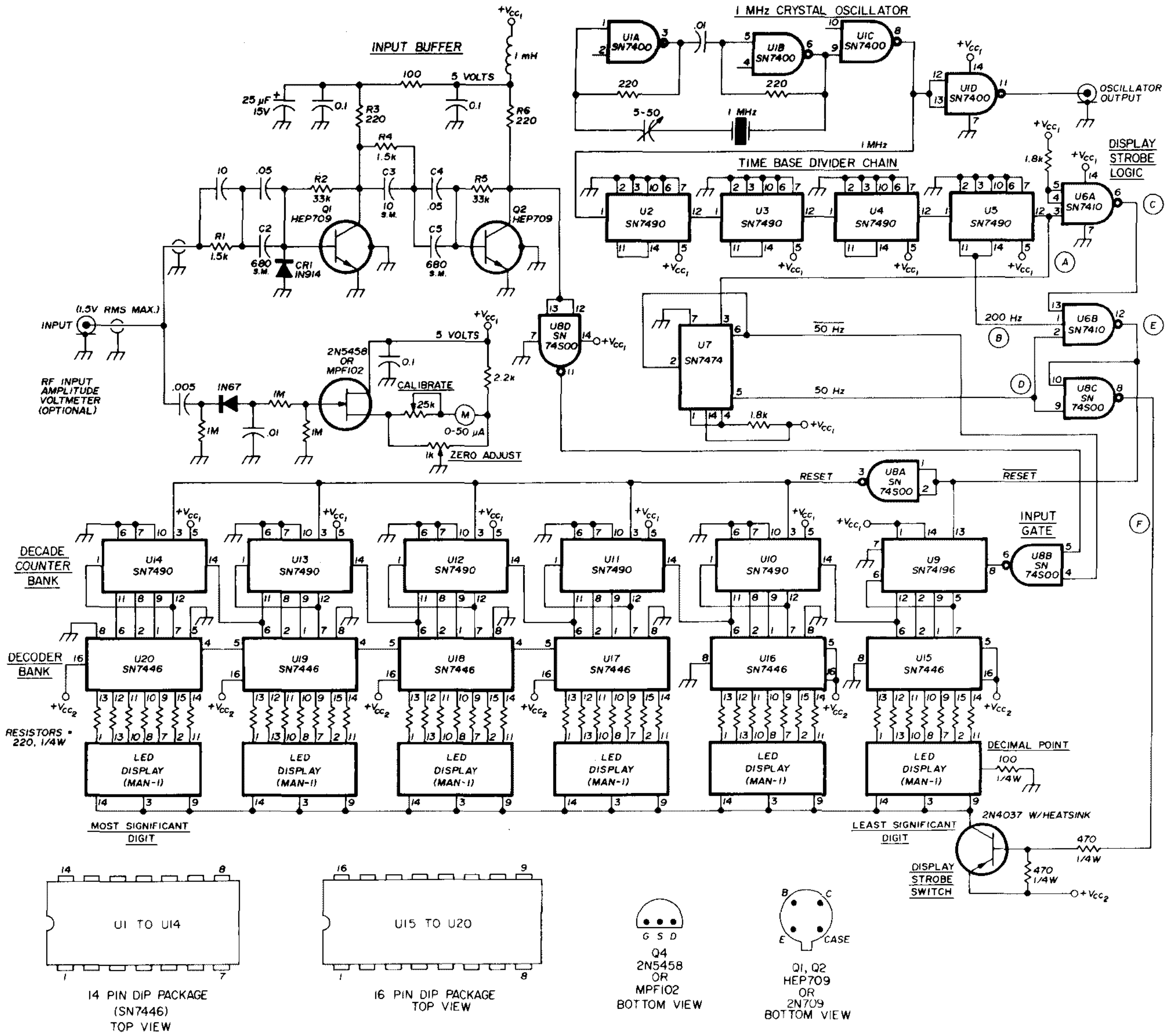


fig. 1. Frequency counter schematic.

(U7) that divides the crystal oscillator down to a frequency of 50 Hz. The flip-flop has two oppositely phased outputs, 50 Hz and $\overline{50\ \mathrm{Hz}}$. The $\overline{50\ \mathrm{Hz}}$ output is 180 degrees out of phase with the 50 Hz output. Each output is a symmetrical square wave that is logic 1 for 10 milliseconds, and logic zero for 10 milliseconds, for a total time period of 20 milliseconds. The $\overline{50\ \mathrm{Hz}}$ from flip-flop U7 controls the input gate (U8B). U8B will only pass the amplified input signal from the unknown source when the $\overline{50\ \mathrm{Hz}}$ at pin 4 of U8B is logic 1. Thus U8B gates the unknown frequency for 10 milliseconds.

The decade counters in the time-base divider chain are connected in a divide-by-5, divide-by-2 configuration. The output frequency of each decade counter is 1/10 the frequency of the input. The output of each decade counter is a symmetrical square wave. The schematic of fig. 2 shows in detail how the decade counter functions when connected in this fashion.

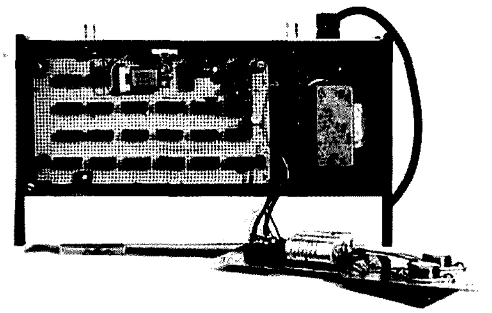
display strobe logic

The display strobe logic (U6, U7, U8) synthesizes the timing sequence for sampling the input frequency, resetting the decade counter bank before each sampling period, and strobing the LED displays once for each completed sampling period. The timing diagram, fig. 3, illustrates the relationship between these signals. "F" is derived from the output of U8C, pin 8. The total time period for F is 20 milliseconds, which is determined by flip-flop U7, as discussed earlier. The duty cycle of F is determined by gating the B, C, D, and E signals together. You'll note that F is high for 11 milliseconds and low for 9 milliseconds, and that the displays are blanked out

during this 11 milliseconds of each sampling cycle. The display strobe switch, Q3, is held in cutoff as long as F is logic 1.

The $\overline{RESET}$ pulse, E, goes low for the first 1 millisecond of the sampling cycle resetting decade counter U9 to the 0000 state. E is inverted by U8A to provide the proper reset signal for the remaining decade counters (U10-U14).

During the 10 milliseconds that follow the trailing edge of E, the input frequency to be counted is registered by the decade counter bank. The LED displays are



Logic circuitry for the 50-MHz frequency counter is built on a section of perf-board. Voltage regulator ICs are mounted on aluminum panel which is sandwiched above the perf board on standoffs.

blanked out during the count-up cycle; otherwise, the displays would show a blur of 8s from the fast count rate. When the sampling has been completed, the input gate is opened, and the decade counter bank no longer receives pulses from the input buffer. At this point in time F goes low; Q3 is switched into saturation, and the LED displays indicate the results accumulated during the sampling period. This process is repeated 50 times per second. Because of the persistence of the human eye, the displays seem to be on continuously. Since the counter gets an update 50 times per second, the counter will follow rapid changes in frequency, such as those encountered when tuning across the band. The counter will update changes in frequency with no apparent time lag.

decade counter bank

The decade counter stages (U9-U14) are cascaded in a manner that allows them to function as a system for counting a series stream of pulses. U9 is the most important link in the counter chain and is an SN74196, a high-speed device capable of performance in the 50-MHz region. The SN7490 decade counters are rated at 15 MHz. Therefore, the frequency range is very dependent upon the input buffer and the SN74196. Since U9 will operate at 50 MHz, the frequency propagated to the next stage will be, at most, 5 MHz. Each succeeding stage will receive decreasing orders of magnitude of the frequency presented to U9.

The SN7490s are connected in a divide-by-2, divide-by-5 format for use in the decade counter bank. Pin 14 is used as the clock input, and the output of the first flip-flop (pin 12) is connected to pin 1 to drive the divide-by-5 portion. The counting function is performed in the binary coded decimal format. Pin 3 is used as the reset input for initializing the decade counters to the 0000 state. A logic 1 at this input will reset the SN7490. When pin 3 is logic 0, the SN7490 advances into each succeeding count state as dictated by the clocking signal at pin 14.

The SN74196, on the other hand, is nothing more than a super-fast SN7490 and operates in much the same manner. The subtle differences are in the pin configuration and the resetting scheme. Unlike the SN7490, the V_{CC} and ground pins on the SN74196 are 14 and 7 respectively; on the SN7490, they are 5 and 10 respectively. Pin 13 on SN74196 is the reset input; a logic 0 as this input will jam the counter into the 0000 state. The counter can only advance when pin 13 is logic 1. This one criterion is opposite that of the SN7490. NAND gate U8A solves this dilemma by providing oppositely phased reset signals for U9 and the remaining counters in the decade counter bank.

decoder bank

U15 through U20 are BCD-to-seven-segment-decoder ICs. These SN7446s translate the BCD information from their respective decade counters to form digits in the seven-segment format. The SN7446s feature leading-zero blanking, which is employed to eliminate any ambiguity caused by one or more zeros preceding the most-significant digit. For example, a frequency of 00142.7 kHz is more recognizable when presented as 142.7 kHz. Blanking out the unnecessary zeros makes the display much easier to read. Special logic is designed into the SN7446 to provide this feature. The ripple blanking logic looks for a logic 0 from the ripple blanking output (RBO) from the next most-significant digit. This condition occurs when the next most-significant digit above that one is also a zero. The ripple blanking signal propagates from the most-significant digit to the least-significant digit desired in the zero blanking scheme.

If you refer to fig. 1, you'll notice that the ripple blanking originates from U20 (the most-significant digit) and is passed down the line to U17. The ripple blanking output (RBO) appears at pin 4 of U20 and is fed to the ripple blanking input (RBI), pin 5 of U19, and so on.

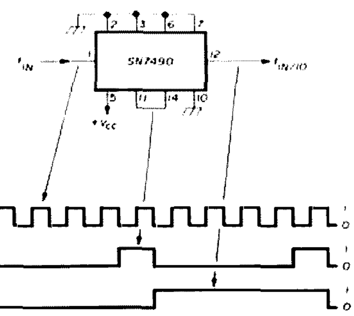


fig. 2. The type SN7490 connected as a symmetrical divide-by-10 counter.

Pin 5 of U20 is grounded since the ripple blanking process originates at U20. The ripple blanking feature can be disabled by simply connecting pin 5 on U20 to $+V_{CC2}$. Pin 5 (RBI) on U15 and U16 are tied to $+V_{CC2}$; therefore, with no signal present at the input gate the counter will display only the least two significant digits as zeros.

Unlike the other ICs in the project, the SN7446s are enclosed in a 16-lead dual inline package. Pin 16 is the $+V_{CC}$ input and pin 8 is ground. The output pins, 9 to 14, are open-collector outputs capable of sinking 20 mA

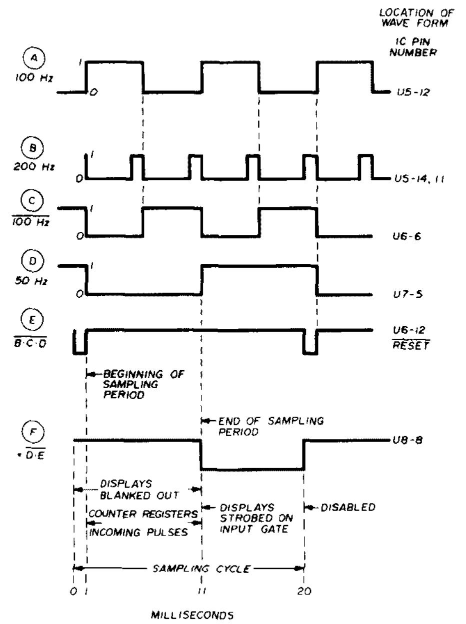


fig. 3. Timing diagram showing display strobe logic, reset, and input gating.

when switched on. The 220-ohm resistors in series with each segment on the LED displays limit the current to a peak value of 15 mA per segment. However, since the displays are strobed on for 9 milliseconds in each 20 millisecond sampling cycle, the average current per segment is (9 milliseconds/20 milliseconds) (15 mA) = 6.75 mA.

led displays

The LED displays used in this project are equivalent to the famous MAN-1. The pinout configuration and schematic are shown in fig. 4. The forward-bias threshold on each segment is slightly more than 1.6 volts. This property alone makes it virtually impossible to check out junction continuity and performance with multimeters equipped with an ohms-scale voltage source of 1.5 volts. The best way to check out the LED displays is to use a 4.5- to 5-volt supply with a series current-limiting resistor of 220 ohms. If purchase of MAN-1s from some of the surplus dealers is contemplated, this setup will prove valuable in judging display performance on a segment-by-segment basis. The displays will appear to be a little dimmer in the finished counter because, as pointed out earlier, the average dc current through the segments is 6.75 mA.

The MAN-1s are common-anode displays. Common-anode displays can only be used with decoder ICs like the SN7446 because of the polarity of the open-collector outputs. Common cathode displays will not work in this project. The MAN-1 display has its segments partitioned into three groups. It is necessary to tie all three *common* elements together (pins 3, 9 and 14) to get all of the segments to light up on command. Litronix Data Lite 707 and the Opcoa SLA-1 are excellent substitutes.

input buffer-counter preamp

The input buffer stage is designed to amplify low-level signals to the amplitude necessary to drive TTL logic circuitry. The transistors chosen for this two-stage amplifier are the HEP 709s by Motorola. The gain-bandwidth product of these devices is 600 MHz, which makes them well suited for this application. Their low saturation voltage ($V_{ce(sat)}$) is on the order of 0.3 volt, low enough to ensure a logic 0 at the input of a TTL device.

Resistor R1 acts as a buffer between the transmission line input and the base circuit of Q1 so that the incoming signal is not clipped or loaded down by the base-emitter junction of Q1. The parallel combinations of C1-C2 and C4-C5 provide coupling from several kHz through the vhf region. Ceramic capacitors become somewhat lossy and inductive at high frequencies, so silver-mica capacitors (C2 and C5) are used to provide additional coupling at the high-frequency end of the counter range.

Diode CR1 is a high-speed switch that protects Q1 from negative-going peaks appearing at the base-emitter junction. Resistor R4 matches the collector circuit of Q1 to the base circuit of Q2, and also contributes to overall amplifier stability. C3 is a 10 pF silver-mica capacitor that compensates for the base-to-emitter capacitance of Q2.

To keep stray capacitance to a minimum, short com-

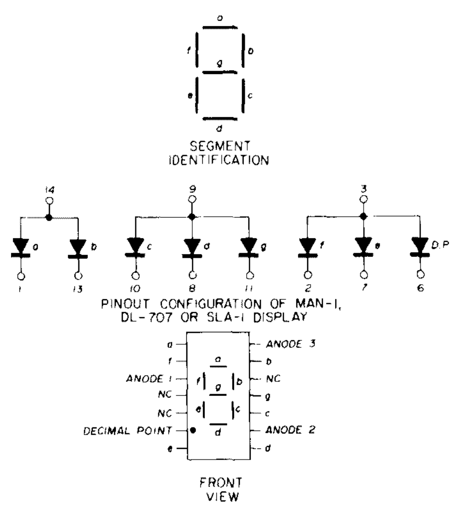


fig. 4. Pinout connections and schematic of the MAN-1 LED display.

ponent lead lengths are of prime importance. The braid of the coax cable should be soldered as close as possible to Q1's emitter to prevent ground-loop problems. Printed-circuit boards with the customary ground planes are not necessary if the layout is as neat and compact as possible. The amplifier should be near U8 since the collector of Q2 drives pins 12 and 13 of U8.

input amplitude voltmeter

Since digital logic has a threshold with respect to triggering levels, a means of monitoring the input signal level is necessary to ensure that the counter chain is receiving enough drive to operate reliably. Insufficient drive level can cause triggering errors, in which case the counter counts only a few pulses that happen to break the threshold level.

This circuit consists of an fet voltmeter equipped with an rf probe. The meter is calibrated to read 5 volts peak-to-peak or 1.78 volts rms full scale. The 25k trimpot in series with the 0-50 microammeter is used to calibrate the circuit. The 1k trimpot provides an electrical zero adjust. Calibration can be done with any high-frequency source of known amplitude. The 1-MHz output at pin 11 of U1 has an amplitude of about 3.6 volts peak-to-peak, which can be used if no other calibrated source is available. Before beginning the calibration, the meter should be both mechanically and electrically zero adjusted. The zero adjust on the front panel of the meter should be checked before you apply supply voltage to the fet voltmeter circuit.

power supply

The power supply, **fig. 5**, is straightforward thanks to the LM309K voltage regulators. Two 5-volt supplies are derived from the 9-volt dc supply. The V_{CC1} supply is connected to the V_{CC} pin of all the ICs except the decoders. The V_{CC2} supply powers the decoder ICs and the LED displays only.

The purpose of splitting up the power supply in this fashion is to divide the current demand of the frequency counter so that the regulators operate well below their maximum ratings. The dual power supply also provides excellent decoupling between the decoders, display switching circuitry, and other parts of the counter logic.

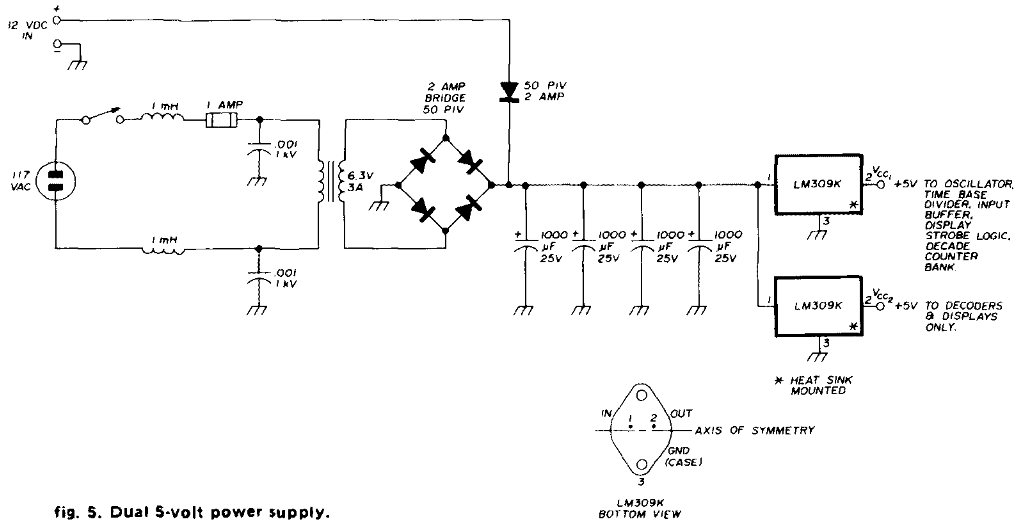


fig. 5. Dual 5-volt power supply.

sensitivity measurements

These measurements were made with a Tektronix 191 constant-amplitude rf generator, a Hewlett-Packard audio oscillator, and a Tektronix 7000 series scope. The following results were observed with a sine wave input. Upper and lower cutoff frequencies of the counter were noted with respect to a given input amplitude. These numbers represent input levels necessary to ensure reliable triggering of the decade counter stages.

amplitude (mV rms)	lower cutoff frequency (kHz)	upper cutoff frequency (MHz)
5 mV	200	10.20
10 mV	150	14.00
15 mV	100	18.15
20 mV	33	23.00
50 mV	20	45.70

The counter works well with signal levels up to 1.5 volts rms (4.5 volts p-p). At greater amplitudes, the base-collector junction of Q1 is forward biased during the positive peaks of the input signal thereby degrading its vhf performance.

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ham radio

antenna and tower restrictions

A complete discussion of deed restrictions, zoning ordinances and building codes, and how they may affect that new antenna system you want to install

When planning to buy or build a new home an amateur's first thoughts are inevitably of antennas. How many towers, and how tall shall they be? He may remember stories of amateurs who have been faced with lawsuits because of deed restrictions or zoning regulations, but rarely is this an overriding consideration in the choice of a house or lot. Let the wife pick out the house first -- he can worry about those things later. This should not be. The dangers to ham operation are real, and unfortunately seem to be getting greater.

No one can assume that he has an inalienable right to do whatever he wishes with his property. Like all of our rights, they are subject to many limitations. If you take the attitude that some vaguely-worded deed restriction or zoning ordinance will be decided in your favor by the courts, you *may* be right. However, it could still cost you thousands of dollars in legal fees to establish your rights, and unless you are inedpendently wealthy and enjoy litigation, it could by a Pyrrhic victory.

I recently bought a lot and built a house, and in the course of doing this learned a great deal about the subject. Because of various restrictions, I rejected lots that were otherwise very desirable from the standpoints of location and price. Eventually I found a lot that was satisfactory from all standpoints, but it was not easy. This article will describe the nature of the problems you may be faced with, and what you should do to minimize your risk.

Perhaps the best way to describe deed restrictions is to demonstrate how they work: The owner of a tract of land wants to subdivide it into building lots and sell them. If the owner of the tract is also a builder, he wants to sell you a house with the lot. It is naturally his desire to get as high a price as possible for his lots, and it is therefore in his interest to impress you with the desirability of living in his development. He wants to convince you that the area is definitely high-grade, and will, furthermore, remain that way and not deteriorate. He drafts a "Declaration of Restrictions." This document generally describes the type of house, garage, etc., you may erect on the property, the minimum setback, type of fence, and other items. The developer submits his plat and restrictions to the local zoning commission or other cognizant authority, and if they comply with local planning and zoning laws, they are approved and recorded. The deed to your property will probably say "Subject to any restrictions of record" or something similar. The restrictions are now legal.

Anyone who buys property in this subdivision is, in effect, signing a contract to abide by these restrictions, and if he violates them he can be sued by any property owner or group of property owners in the subdivision.

Of course, there is no certainty that you will be sued if you violate the restrictions. But you are certainly subject to lawsuits. If the development is new, the developer himself may sue, since he may feel that the presence of a 70-foot tower makes it more difficult for him to sell his remaining lots. But even after the subdivision is all sold out, at which time the original developer rapidly loses interest in the character of the neighborhood, any property owner can sue if he finds your tower objectionable, and, depending on the exact wording of the restrictions, would probably have a good chance of winning. The result would be a court order for you to remove your tower.

In my search for a lot, I accumulated quite a collection of sets of restrictions. Every one of them, to a greater or lesser degree, implied restrictions on the erection of antennas although the wording varied. In fact, most of them made no mention of antennas as such. These specified in detail what you *could* put on the property – and an amateur antenna was *not* one of the permitted things. Typical wording was, "No structure other

Harry R. Hyder, W7IV, 9842 North 57th Street, Phoenix, Arizona 85253

than a single-family dwelling . . . garage, swimming pool, etc . . . shall be erected or permitted to remain on this property.

One local amateur was sued by a developer for violation of a restriction like that. He won his case because his tower was mounted on top of his carport, and the court ruled it was part of the house and not a separate structure. Presumably if the tower has been mounted on the ground apart from his house, he might have lost. This victory cost the ham $2500 in legal fees. Some victory.

Other restrictions specify a maximum height above ground or roof level that no structure can exceed; 35 feet (11 meters) above ground level or 3 feet (1 meter) above roof level are common values.

Some restrictions do mention antennas. One actually specifically permitted amateur radio antennas but said that the towers must be of the retractable type and lowered when not in use. One planned community in the Phoenix area bans all outdoor antennas including TV antennas (this community has a master antenna and cable distribution system).

Another common type of restriction states that anything erected on the property must be approved by the "Architectural Committee" of the development.

When considering lots, I always asked the agent whether there were any restrictions on the erection of amateur antennas, and was usually assured that there were none. This frequently turned out to be not quite true. Very rarely, in fact, did the agent have a knowledge of what the restrictions actually said. Ask the agent to get a copy of the restrictions, which he can easily do, and read them yourself. You can get copies yourself as most of the large title insurance companies have this information on file. Alternately, you can get copies at the County Recorder's office (to be valid, the restrictions must be a matter of record).

All of this sounds pretty discouraging, but there are subdivisions that do not have any restrictions. This is more likely to be true of older areas since it has been only relatively recently that restrictions of this sort have become widespread. There are also many odd pieces of property that have never been part of subdivisions. If you look hard enough, you can find a suitable house or lot that has no restrictions.

In any event, you should have your attorney insert a clause in the sales contract that your money will be returned if any restrictions on the erection of ham antennas are found to exist.

The restrictive covenants, zoning ordinances and building codes which affect amateur radio antennas vary widely from one locale to another, and much of the law is case law which varies greatly from state to state. Some states, for example, have given substantial weight to aesthetics while other states have held that aesthetics cannot be considered at all. And, although the law on restrictive covenents is much more uniform from state to state, there are conflicting opinions in amateur radio cases. Furthermore, the application of common and statutory laws on nuisances is a rather new development which is certain to become more of a problem in the years ahead.

This article is based on the author's experiences in the Phoenix, Arizona, area, so the restrictive covenants, zoning ordinances and building codes may be considerably different then in your own area. Nevertheless, his basic guidelines are applicable in practically every case. Following these guidelines may not keep you completely out of trouble (witness the nuisance cases of WØMYN and W2OVC), but they should give you a foot up on the problem.

Editor

zoning ordinances

Unlike deed restrictions, which are in the nature of private contracts, zoning regulations are a matter of law. Zoning is an attempt by a municipality or county to control the usage of land within its boundaries for the purpose of orderly growth and development. Sections of the incorporated area are designated residential, commercial, industrial, etc., and within these classifications are sub-classifications.

The various types of residential zoning control the number of residences per acre, whether single-family or apartment buildings, maximum height, setback from the street and property line, street and utility easements, and similar matters. Depending on the exact wording, zoning regulations can imply prohibition of the erection of antenna towers, or can expressly forbid them. Some even expressly permit them. The zoning regulations of Scottsdale, Arizona, for example, specifically permit antenna towers up to 70 feet (21 meters).

In contrast, Paradise Valley, a bedroom community adjoining Phoenix, forbids *all* towers. An amusing sidelight to this is that Paradise Valley's most distinguished citizen is Senator Barry Godlwater, K7UGA. Senator Goldwater's home is equipped with two tall towers mounting quite an array of beams, including a very impressive log-periodic.

The Phoenix Zoning Ordinance controls building height, but a paragraph specifically excludes antennas, flagpoles, water towers, etc., from the height restrictions.

In general, amateurs are in less danger from zoning ordinances than from deed restrictions. One reason is the natural slowness of democratic governments to react except in the face of political pressure. Another is the fact that most municipal or county attorney's offices are very understaffed, and are not anxious to undertake such suits, which do not have the glamour of, say, criminal prosecutions. Nevertheless, city and county attorneys are usually elected officials, and if one of your neighbors objects to your antenna, and he is politically well-connected, you could be in for trouble.

The amateur does have one thing going for him. There seems to be an unofficial legal principle that says what others have done in the past without legal interference, you can do too. If there are a number of amateurs in your city who have towers and have never been threatened with legal action, regardless of the exact wording of the zoning regulations, you are probably on safe ground.

In any event, it is a good idea to become familiar with your local zoning regulations. These can be obtained from your local planning and zoning commission, usually located at city hall or nearby. The complete Phoenix Zoning Ordinance, a sizable book, costs $5.00, and by paying an additional $5.00, you can be placed permanently on the mailing list for changes and additions. Other cities probably have similar arrangements.

building codes

Building codes are designed to protect the health and safety of the citizens of a political division. Antenna towers come into this because an improperly designed or installed tower could collapse and cause damage to life or property. I have personally seen antennas that seemed to stay up by sheer faith, and it seems reasonable that anyone erecting a tower should be able to demonstrate that it will not be a hazard to himself or his neighbors.

Relatively few amateurs apply for building permits for towers. I strongly recommend it. In principle, at least, if you do not get a permit with its attendant inspections, you could be forced to take your tower down. It's difficult to say just how likely this is to happen as it is highly dependent on your local administration, but it can and has happened. In any event a building permit is a nice piece of insurance against that possibility.

In some localities obtaining a building permit is mere formality. Some areas even have special provisions in their local codes for amateur radio towers. In others it is more difficult. A typical requirement would be a set of plans and stress calculations approved by a registered professional engineer. First find out from your local building inspection department what is required, and then attempt to supply it.

One problem that you may run into is that the personnel in your local inspection department have never been asked to issue a permit for a tower before, and like true bureaucrats, assume that since they have never done it before, it must be illegal. Don't stop there.

"They're recalling your outside antenna."

An experience of mine is enlightening. When I was interested in buying a piece of property in a local community, I went to the inspection office and asked the man behind the counter how to apply for a permit for a 50-foot (15-meter) antenna tower. He informed me very positively that such a tower was not permitted and he could not issue me a permit. I then asked him why the local zoning ordinances permitted antenna towers up to 70 feet (21 meters) if they were illegal (I had already checked this). I then showed him the wording of the ordinance, and he confessed that he had never seen that paragraph. He allowed as how he probably could issue a permit, but would have to check with the city engineer about the actual requirements. I did not pursue the matter at that time, and I eventually bought my lot elsewhere, but an amateur friend of mine subsequently received a permit for a 60-foot (18-meter) tower from the same office merely by supplying a set of the tower *manufacturer's plans and specifications.*

Don't stop at the first "No." Building inspection departments are bound by law and cannot act arbitrarily. If the zoning laws do not prohibit towers and if you can demonstrate that the tower design and installation are sound, they are bound to issue you a permit. The cost is usually about five dollars.

It would be a good idea to read the sections of your local building code dealing with towers. You can do this at the inspection office. If you run into any problems it would be a good idea to request a personal talk with the city or county engineer. They are usually pretty reasonable.

In the city of Phoenix the law requires that the tower stress calculations be checked by a registered professional engineer in the *State of Arizona* and that he supply a letter saying that he has done this, duly stamped with his seal. I don't know how widespread this requirement is, but it could mean a fee of $50 or more. In my own case, the tower I was planning to erect had already been approved for a previous applicant. Some areas have approvals for specific makes and models of towers on record; this constitutes a sort of type acceptance. It would be a good idea to find out which types have previously been approved, and if one of these suits your requirements, getting a permit should present no problems.

Another thing to do is to find a local amateur who has successfully obtained a building permit and find out what procedure he followed. When dealing with the law, precedent is highly important.

conclusions

While it may be troublesome and frustrating to run the gauntlet of deed restrictions, zoning ordinances and building permits, it can be done — the most important ingredient is persistance and it is worth it. Amateurs have been ordered by the courts to remove their antennas for violation of all of these, and it gives you a comfortable feeling to know that you are completely protected.

ham radio

diode detectors

A comparison of the operating characteristics of various diode detectors and how they can be improved through modern circuitry

From the earliest days of radio the subject that received the most attention of radio amateurs (first unlicensed, and later with amateur calls) was the detector. The antenna-ground system, although it allowed for a good deal of innovation, was generally size-limited by the amateur's real estate or by the basic laws of physics. Also, antennas are fun to work on only in decent weather; little antenna work is done during the winter months. The transmitter was also straightforward: you simply bought as large a transformer as you could afford. The spark gap and its coupling to the antenna-ground system were relatively simple.

The detector, however, didn't cost much and could be worked on at any time so thousands of experimenters tinkered away their winter evenings trying to improve their detectors. Eventually they had enough success that the detectors became known as "receivers." Another nice feature of experimenting with detectors was that you could receive signals to "get a foot in the door" of radio, even if you had no transmitter. All sorts of devices were tried by these early experimenters: flame ionization detectors, coherers, electrolytic detectors, thermoelectric detectors, magnetic hysteresis detectors, crystal detectors and the early Fleming valve (vacuum diode).

Of all the early detectors the crystal type received the most widespread usage and "crystal set" eventually became a household word. Various types of mineral and man-made crystals such as galena, silicon, perikon (copper pyrites and zincite), molybdenite and carborundum were used. **Fig. 1** shows a simple crystal set using one of the crystals of the period. This same circuit could still be used today, but a modern signal diode would be used in place of the crystal and catwhisker.

Galena (lead sulphide), an important *lead* ore found here and in Europe, was the most popular of the crystals used in the early days of radio because it was the most sensitive. Steel galena, so called because it resembles a piece of broken steel rod, contains a small percentage of silver and, although not quite as sensitive as plain galena, became popular in later years because it was somewhat *easier to adjust.*

The crystals used as radio detectors were mounted in clips, held in tin-foil cups, floated in mercury, or more commonly, mounted in a small "pill" of a low-melting-point alloy. (Some experimenters who tried to mold their own crystal holders used a too-hot mixture of lead, only to discover that the heat destroyed the sensitivity of the galena.) The catwhisker, a length of fine, stiff wire,* was moved about the surface of the crystal until an "active" spot was found. This metal-to-semicon-

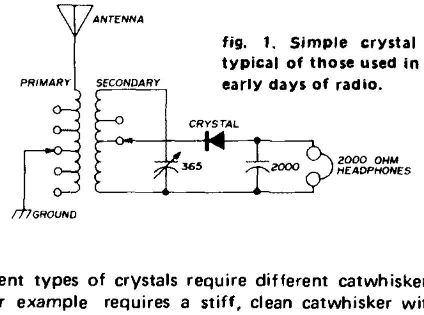


fig. 1. Simple crystal set typical of those used in the early days of radio.

*Different types of crystals require different catwhiskers. Galena, for example requires a stiff, clean catwhisker with very little pressure. Plated copper is best, with brass and platinum running a close second. If you use a *steel* galena crystal, however, a German-silver catwhisker is best. For silicon crystals, tungsten catwhiskers are preferred although molydenum is sometimes used. Chromium or steel are recommended as catwhiskers for carborundum crystals (which also require a bias battery), while many different metals have been used successfully with molybdenite crystals.

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

ductor interface is similar in many ways to the point-contact diodes of today. A typical galena-catwhisker assembly is shown in fig. 2.

The success of the vacuum tube, first as a diode (Fleming valve) and later as a triode (Deforest Audion) eclipsed the crystal detector commercially after about 1921. Crystal sets continued to be used by experimenters, however, as they still are today. Also, work continued in the laboratory on silicon crystal detectors as power detectors for microwave measurements.[1] This laboratory experimentation was greatly refined and expanded during World War II as engineers tried to solve the microwave radar mixer problem. This concentrated research effort on crystal detectors eventually led to a huge body of knowledge on basic silicon and germanium crystal physics and how various impurities affect the semiconducting properties of these materials.

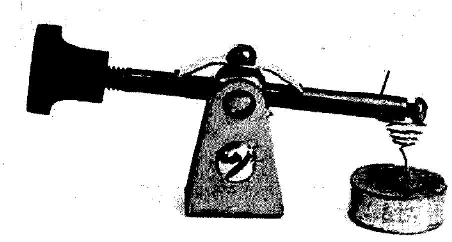
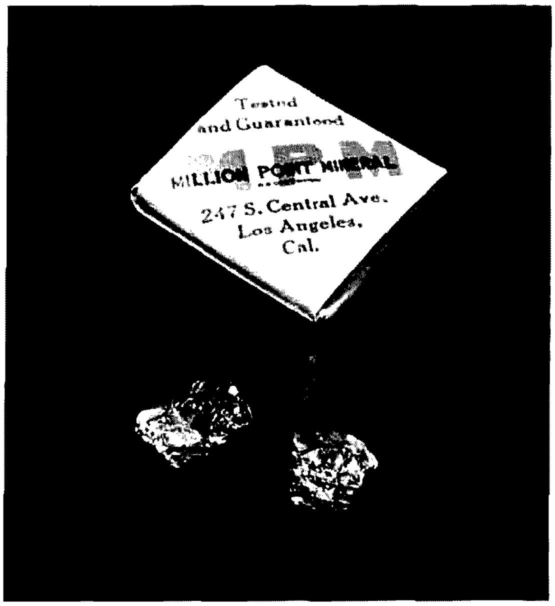


fig. 2. Commercial galena-catwhisker assembly. These units can still be obtained at some radio distributors.*



Steel galena crystal offered to amateurs in the 1920s.

The high-inverse voltage germanium point-contact diode came directly out of these wartime research efforts, and most of the other semiconductor developments we know today came indirectly from this same research. The twenty-eight volume MIT Radiation Laboratory Series includes one whole volume which describes the semiconductor diode developments for radar usage which occurred during this period.[2]

Based upon the research done during the war, germanium point-contact diodes became available to industry in the late 1940s. The 1N34 was offered by Sylvania, immediately became popular with experimenters, and started showing up in everything from absorption wavemeters to a-m speech clippers.[3]

The alloyed-junction silicon and germanium diodes came along in the early 1950s.[4,5] The germanium junction diode achieved some degree of popularity as a rectifier (1N91 – 1N93) but is considered obsolete today while the silicon junction diode came into its own *both* as a signal diode and power rectifier. An example of an early alloy-junction rectifier is the 1N536-1N540 family; the 1N482-1N485 family are typical early alloy-junction signal diodes.

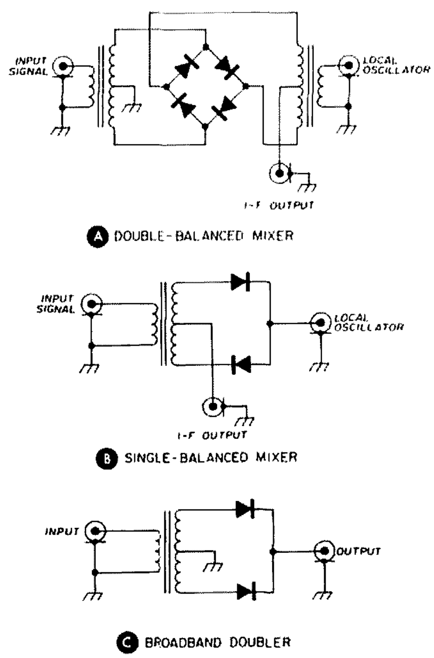


fig. 3. Three diode circuits which use Schottky diodes. Shown in (A) is a typical double-balanced mixer. A single-balanced mixer is shown in (B) while (C) shows a broadband doubler circuit.

*Mounted galena crystals, crystal stands and catwhisker assemblies are available from Modern Radio Laboratories, Post Box 1477, Garden Grove, California 92642. Their catalog, available for 25¢, also lists an assortment of crystal-set kits and other hard-to-find items such as carborundum detectors and coil sliders.

While most modern silicon junction rectifiers are still made by the alloy process (1N4001-1N4007, for example), newer silicon junction signal diodes are usually made by the planar epitaxial process (1N4454, for example). If you insist on a germanium junction signal diode, the base-emitter elements of a germanium junction transistor (2N404 or 2N5043) could be used.

There have been many other types of diodes developed since the silicon and germanium types discussed above. Tunnel, PIN, step-recovery, varactor, zener,

Virtually all Schottky diodes are silicon types, and their advantage over point-contact types is that their characteristics are closely matched and *stable*. This stability quality is quite important because it allows the close *matching* of diode pairs or quads which make it possible to build really good double-balanced mixers. The double-balanced mixer and its related single-balanced mixer and broadband doubler have made an enormous impact on modern vhf, uhf, and microwave systems.

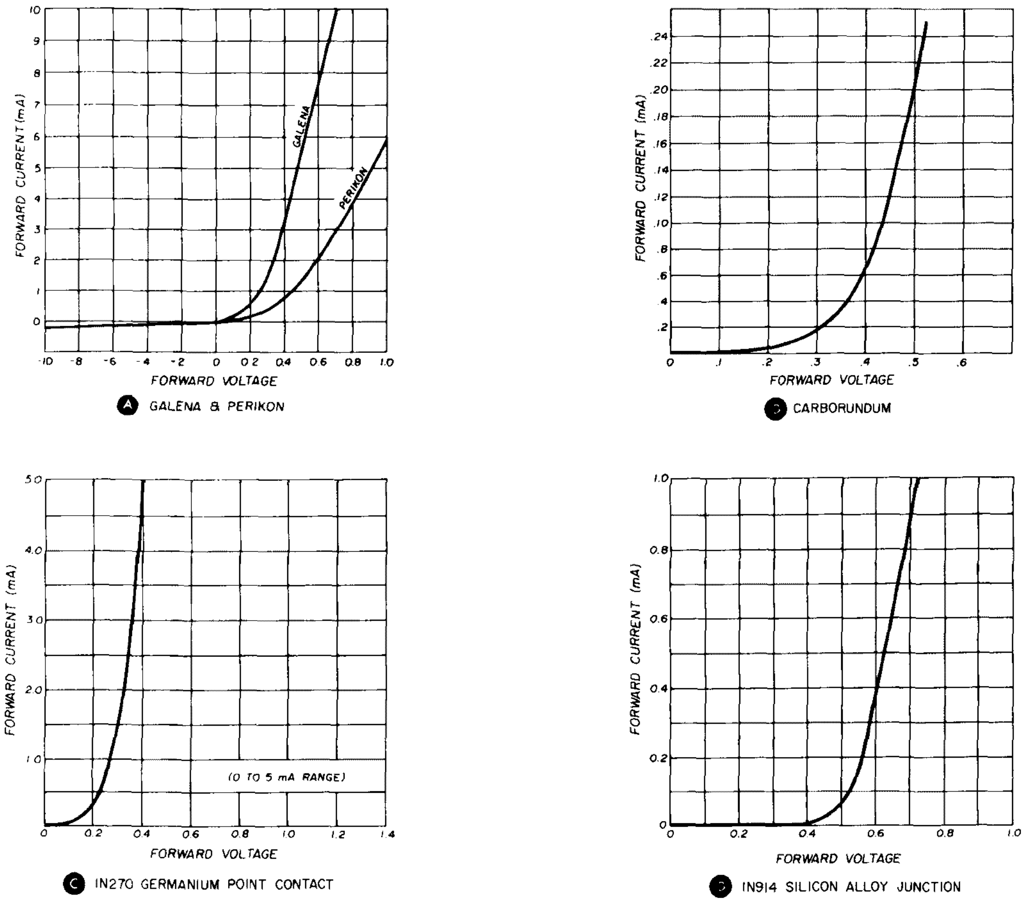


fig. 4. Forward conduction characteristics of several different diode types including galena and perikon (A), carborundum (B), germanium point contact (C) and silicon alloy junction (D).

Gunn, IMPATT and TRAPATT are some of these special types, but are not, in general, used as detectors of the common, rectifying type.

One newer type of diode, the Schottky-barrier or hot-carrier diode, has characteristics similar to the point-contact device. Like the point-contact diode, the Schottky diode uses a metal-semiconductor junction; in the Schottky diode, however, the metal is deposited on the semiconductor by sputtering in a vacuum. Examples of the Schottky diode are the Hewlett Packard 5082-2800 and the Motorola MBD501.

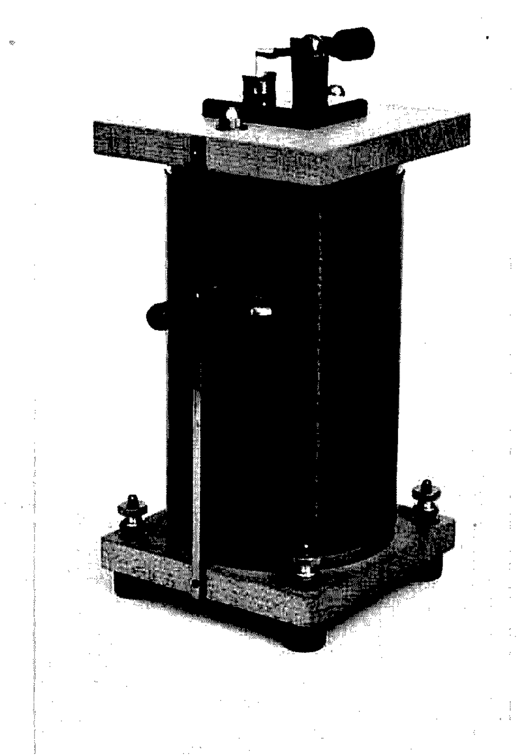
Three basic diode circuits which use matched Schottky diodes are shown in **fig. 3**. Note that the transformers in **fig. 3** are usually built with a few turns of wire on ferrite toroids, so the circuits are often useful over a three-decade frequency range (200 kHz to 200 MHz is common).

The forward conduction characteristics presented in **fig. 4** will give you an idea how some of the various semiconductor diodes compare. The curves in **fig. 4A** for galena and perikon are from reference **6**, the carborundum curve (**fig. 4B**) is from reference **7**, while the for-

ward characteristics for the 1N270 (fig. 4C) and 1N914 (fig. 4D) were taken from the data sheets of currently manufactured diodes. As can be seen, the forward current characteristic of any diode semiconductor diode is far from linear.

diode detectors

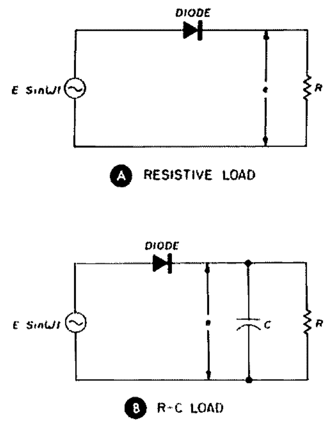
An early article by Colebrook exhaustively describes crystal rectifiers and their use as detectors of radio signals.[6] Although the author had only early galena and perikon diodes to work with, his mathematical analysis and conclusions are as fresh today as they were when written in 1925. The basic diode detector shown in fig. 5 is the same as that used in reference 6.



Crystal receiver of the type used by amateurs sixty years ago. Station was tuned in by moving the slider along the tuning coil. Selectivity was very poor, but there were few stations on the air and a local spark transmitter wiped out the DX anyway.

Fig. 5A shows a resistive load while fig. 5B shows the more usual case where there is a capacitor in parallel with the load resistor (this increased detector efficiency). The capacitor should present a low impedance at the carrier frequency (as compared to resistor R), and a high impedance at the modulation frequencies. In the circuit of fig. 1 a 0.002 μF capacitor is placed in parallel with a set of 2000-ohm headphones. At 1 MHz (the center of the broadcast band, for which this crystal set

fig. 5. Basic diode detectors showing simple resistive load (A) and more usual case where a capacitor is placed in parallel with the load resistor (B).



was designed) a 0.002 μF capacitor has about 80 ohms reactance. At 1000 Hz, a typical audio test frequency, the reactance of this same capacitor is nearly 80 kilohms.

The circuit shown in fig. 6 was built to demonstrate how a shunt capacitor increases efficiency, and also to show how several common diode types compare. The 51-ohm resistor at the input simply terminates the amplitude-modulated signal generator. The LM318H IC and associated capacitors and resistors comprise a lowpass filter with a cutoff frequency at 2000 Hz. In this circuit the 2200-ohm resistor, R1, is the detector load and the 510-ohm resistor at the output of U1 is to prevent oscillation of the op amp when using a length of coax to the vtvm. The 0.002 μF capacitor can be switched in or out; the results are shown in fig. 7. For higher input signal levels the capacitor increases the output audio voltage level by 8 to 10 dB (enhancement with the 6AL5 vacuum tube diode is even more marked at some input levels).

Note that since the plots of input rf level vs audio output level presented in fig. 7 are on log-log coordinates (since the abscissa and ordinate are both in dB) two straight lines may be drawn on the plots, one representing a linear relation and the other a square-law relation. For large input signals, say above -20 dBm, all the detectors approach a linear slope. It should also be noted that

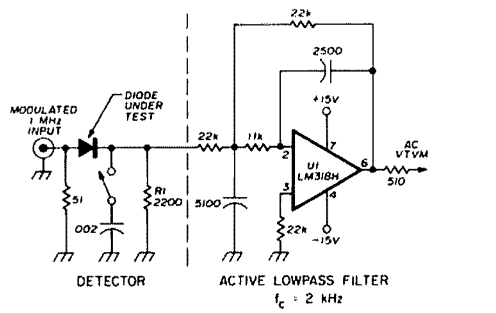


fig. 6. Circuit for testing the operation of various diode detectors. Test results are plotted in fig. 7.

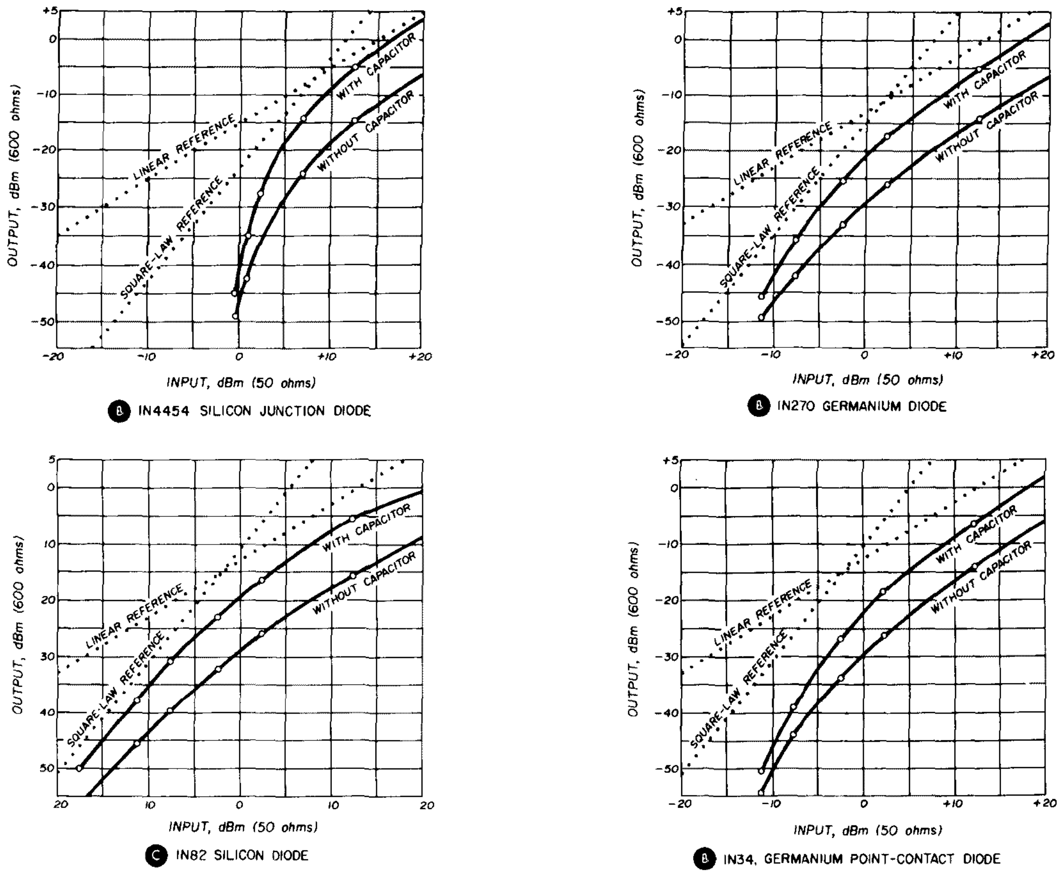


fig. 7. Rf input level vs audio output level for various types of diodes, with and without parallel load capacitor. Linear and square-law references are shown for comparison.

at input levels below −20 dBm a square-law or larger exponent relationship is usual. The point is that although some diodes are more nearly linear over a larger range of input voltage than others, none of them could remotely be considered as linear detectors at input signal levels below −20 dBm. A 6AL5 detector circuit with capacitor, comes closest, perhaps, to the textbook explanation that "diodes are square-law devices for small signals and linear devices for larger signals."

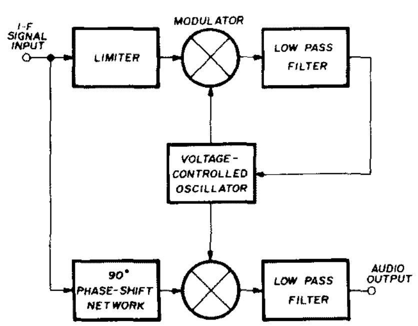


fig. 8. Using a phase-locked loop as an a-m detector (see text).

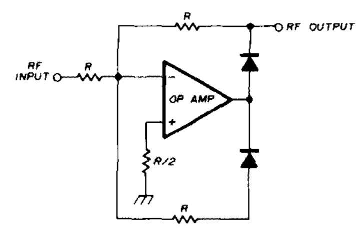
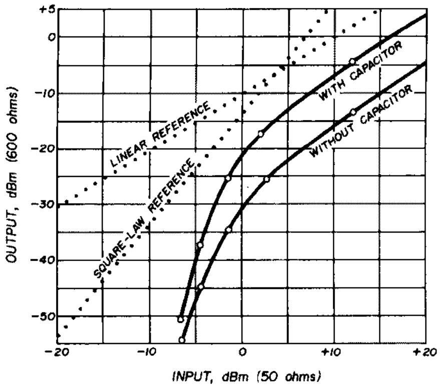
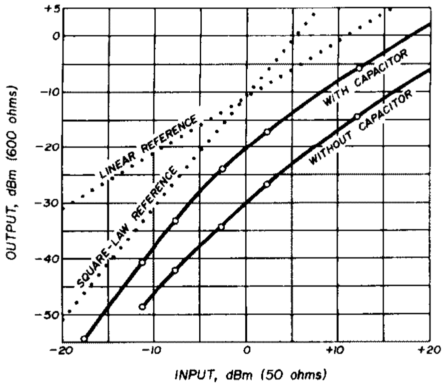


fig. 9. Precision half-wave detector using semiconductor diodes and an operational amplifier.

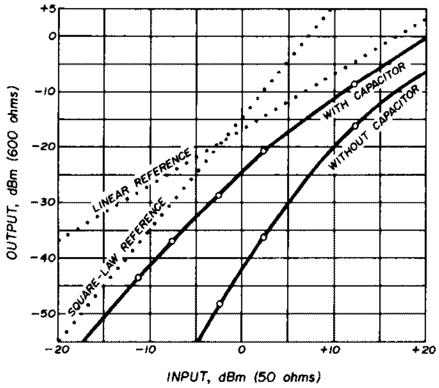
In a receiver, operation of the diode detector in its square-law region means that for every 10 dB weaker a signal may be, the output is 20 dB down. This is clearly not a good way to operate. Not only does it waste stage gain, it also degrades the signal-to-noise ratio. To avoid the square-law region most receivers use enough rf and i-f gain to keep the input voltage level to the diode detector up in the region where it behaves linearly. Unless agc is used this usually means that the last i-f amplifier must be a small transmitting tube or other large-signal device if reasonable dynamic range is to be achieved.



E H-P 5082-2800 SCHOTTKY DIODE



F GALENA-CATSWHISKER



G 6AL5 VACUUM DIODE

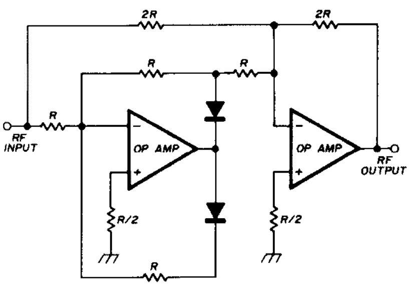


fig. 10. Precision full-wave detector using semiconductor diodes and two operational amplifiers.

linear diode detectors

Few commercial receivers bother with such luxuries as large-signal capability in the last i-f stage – they either rely on agc or *accept* detector non-linearity. Fortunately, the modern extensive use of ssb on the high-frequency bands has forced receiver manufacturers to use the inherently more linear product detector. The linearity of product detectors, which are essentially mixers with an audio output, is due to the fact that the oscillators which drive them completely control the conduction of the nonlinear devices used as mixers.

The product detector does not make a very satisfactory detector for a-m because the bfo never quite matches the receiver carrier frequency. This results in a beat note being present in the audio output unless a phase-locked loop is used to synchronize the bfo to the

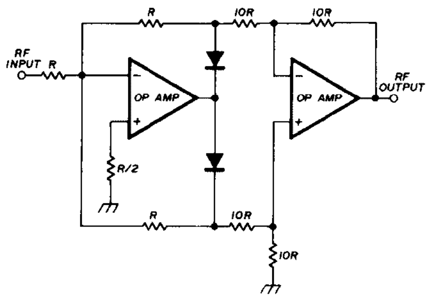


fig. 11. Improved version of the full-wave detector provides better linearity than previous circuits.

received carrier frequency. This phase-locked loop form of a-m detection is shown in fig. 8; with the modern phase-locked loop ICs that are now available the circuit is not unreasonably complex.

Another technique for linearizing an a-m detector involves the use of an operational amplifier. Although this

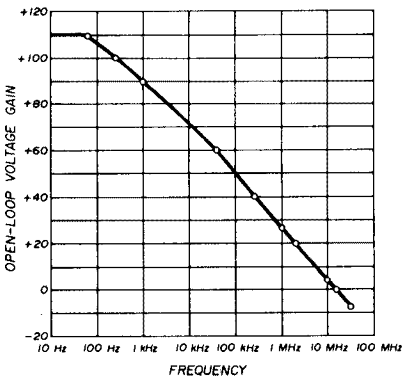


fig. 12. Open-loop gain vs frequency of the LM318H operational amplifier IC.

technique has been around for quite some time, it has only recently become practical with the availability of low cost, high-frequency IC op amps.[8,9] **Fig. 9** shows the basic precision diode detector using an op amp. **Fig. 10** shows a full-wave version of the detector (from reference **9**). The full-wave version has the disadvantage that the delay for positive input signals, which are inverted and amplified two times, is twice that for composite signals. Because of the delay difference, the signals don't subtract in phase, so high-frequency performance suffers. **Fig. 11** shows a precision full-wave diode detector, attributed to Dr. Nick Cianos of SRI, that solves the problem.

only expect about **24** dB improvement in linearity at **1** MHz. To check this I built the circuit shown in **fig. 13.** The test results are plotted in **fig. 14.** The improvement at low input signal level linearity is quite apparent. The principle of op amp/diode detection is used in the National **LM372**, an IC that combines the functions of i-f amplifier and a-m detector.

summary

The diode detector has been the standard a-m detector almost since radio began. Today we essentially have only two choices of semiconductor material: germanium and silicon. Germanium has lower offset voltage while silicon has the benefit of improved technological processing. A semiconductor diode of either type, used in combination with a modern IC op amp, can greatly improve the linear dynamic range of the detector. When used as an integral part of an IC the silicon diode holds great promise in the future.

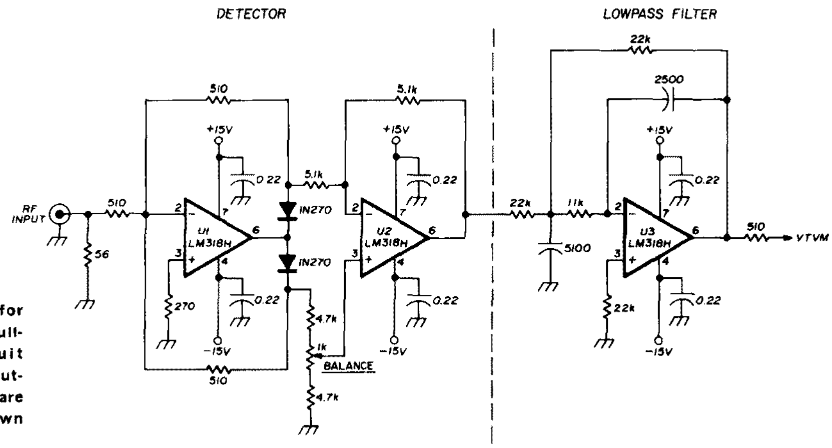


fig. 13. Test setup for checking the improved full-wave detector circuit shown in fig. 11. Input-output characteristics are extremely linear as shown in fig. 14.

Diode detectors which use an op amp in the circuit reduce the input voltage at which the transfer curve (input to output relation) becomes non-linear by a factor equal to the open-loop gain of the op amp. Since op amp voltage gains can be more than 100 dB at the lower frequencies this can make a significant difference in detector performance.

The performance curve of a good monolithic IC op amp (National **LM318H**) is shown in **fig. 12.** Note that the open-loop gain drops as frequency increases so you can

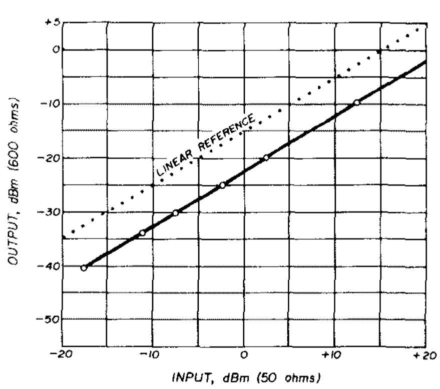


fig. 14. Input-output characteristics of the improved full-wave detector circuit shown in fig. 13 coincides very closely to linear reference.

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ham radio

MICROPROCESSORS

A discussion of microprocessors and how they fit into the scheme of computers and controllers that exist today

By now, most amateurs are aware of the fact that a revolution is occurring in the electronics industry: *microprocessors*. If you had held stock in companies that manufacture microprocessors, this fact would have become quite apparent after RCA's misinterpreted announcement several months ago that microprocessors will soon be incorporated into U.S. automobiles. Rather than rehash an electronics revolution after it is over, we believe that it would be fun to jump into the middle of the one that is occurring at this moment and closely observe events that will have a profound influence on electronic measurement, laboratory instrumentation and amateur station control. Therefore, over the next few months this column will be devoted to the subject of microcomputers: what they are, how they operate, and what they can and cannot do for the electronic experimenter, engineer or laboratory scientist.

We shall use microprocessor operation and interfacing as a vehicle to probe more deeply into the detailed concepts and techniques of computer interfacing. Please keep in mind that the microprocessor, when complemented by memory, buffers, and input/output (I/O) devices, is as much a computer as its larger and usually faster rivals, the minicomputers and full-size computers. By learning how to interface a microprocessor, you will simultaneously learn the concepts of how to interface a minicomputer or full size computer. The use of interrupts, device selects, software generated strobes, timing loops, and the like are common to all.

To gain full value from some of our forthcoming columns, it would be beneficial to have an understanding of the basic principles of digital electronics. Some very important terms and concepts that you should master include the following: gate, logic element, counter, gated counter, monostable, enable, disable, inhibit, strobe, decoder, multiplexer, demultiplexer, timer, clock pulse,

By David G. Larsen, WB4HYJ, Peter R. Rony and Jonathan A. Titus.

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

positive edge, negative edge, flip-flop, latch, bus, Tri-State™, shift register, dynamic RAM, static RAM, ROM, programmable ROM, up/down counter, AND, OR, NAND, NOR, exclusive OR, arithmetic element, and more. Our pair of books on digital electronics, *Bugbooks I and II. Logic & Memory Experiments Using TTL Integrated Circuits*, will bring you to the level of understanding in digital electronics required to interface microcomputers; other digital books, such as the pair marketed by Hewlett-Packard in conjunction with their logic lab, will also help you develop the skills that you will need. Digital electronics is a rapidly expanding field, and new texts and reference manuals are appearing at the rate of one every several weeks.

As we currently envision them, future columns will offer a tutorial on the operation and interfacing of a very popular microprocessor, the Intel 8080 8-bit microprocessor, which can perform a simple logic or arithmetic instruction in only 2 μsec and can directly address 65,536 different memory locations, each containing eight bits of data. Originally priced at $360 in quantities of one, you can purchase an 8080 now for about $50 from selected supply sources and its cost will be no more than $5 two or three years from today. The 8080 has some important rivals, e.g., the Motorola 6800 and the Fairchild F8, but it is a worthy selection nevertheless. Each microprocessor has its special features. However, the general concepts developed in this column will be applicable to any microcomputer system.

Standing alone, a microprocessor chip can do nothing. It functions only in the context of a microcomputer system, in which appropriate integrated-circuit chips are incorporated to complement the basic function of the microprocessor (μP): to serve as a central processing unit (CPU) in which logic and arithmetic operations and data transfers between register, memory, and the outside world are performed. In some columns, we will need to focus upon a specific microcomputer system. For this purpose, we have chosen a new system that is specially designed to instruct individuals in all of the details of microprocessor operation and interfacing: the Mark 80 microcomputer (**fig. 1**). This particular system, shown with 4k of solid-state memory and a control panel, is built around the Intel 8080 microprocessor chip. Except for a power supply, it is completely operational. The system is bus structured and has all important inputs and outputs connected to a solderless breadboarding socket, permitting interfacing concepts to be learned, tested, and breadboarded into a digital circuit of your own design.

microprocessors: where do they fit?

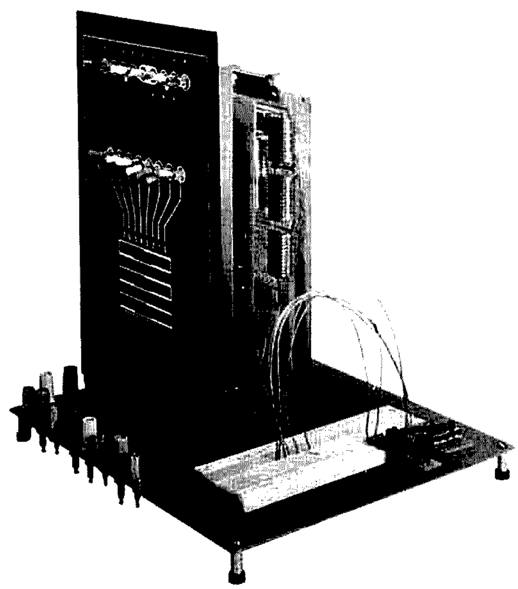
We would first like to discuss what a microprocessor is and how it fits into the general scheme of computers and controllers that exist today. Eadie, in his book, *Introduction to the Basic Computer*, has defined the term *data processor* as "a digital device that processes data. It may be a computer, but in a larger sense it may gather, distribute, digest, analyze, and perform other organization or smoothing operations on data. These operations, then, are not necessarily computational. Data processor is a more inclusive term than computer."[2]

A *microprocessor* is a single integrated-circuit chip that contains at least 75 percent of the power of a computer. It usually cannot do anything without the aid of support chips and memory, however, and therefore can be distinguished from a *microcomputer*, which is a full operational system based upon a microprocessor chip that contains memory, latches, counters, input/output devices, buffers, and a power supply in addition to the microprocessor chip. A microcomputer may be a "black box" with only a single switch: *operate/reset*. The 8080 microprocessor, a 40-pin LSI chip, is shown in **fig. 2**. A typical system based upon this chip is shown in **fig. 3**; the 8080 chip is located on the CPU board on the left.

A microcomputer possesses all of the minimum requirements of a computer. For example:

It can input and output data, which is usually in the form of digital electronic signals. Common I/O devices include teleprinters, CRT displays, paper

fig. 1. The Mark 80 microcomputer system.



tape readers, floppy disks, magnetic tapes, cassette tapes, laboratory instruments, and process control devices.

It contains an arithmetic/logic unit (ALU) that can perform arithmetic and/or logic operations such as add, subtract, compare, rotate left, rotate right, AND, OR, negation, and exclusive OR.

It contains a minimum amount of "fast" memory such as RAM, ROM, PROM, or core, but usually not cards or paper tape, in which data and program instructions are stored. The data and instructions are stored as 4-bit, 8-bit, 12-bit, or 16-bit words.

It is programmable. The data and program instructions can be arranged in any sequence desired, in contrast to the programmable calculator, in which the precise manner that a keyboard function is executed cannot be changed by the operator.

It is fast, with an ability to execute a simple instruction in ten microseconds or less. All existing microcomputers are digital and TTL compatible, where logic 0 corresponds to ground potential and logic 1 corresponds to +5 volts.

There appears to be some misunderstanding concerning the role of current microprocessors and microcomputers relative to other types of computers. The temptation is great to order a modest microcomputer system and then to surround it with $5000 worth of I/O devices such as floppy disks and line printers. At this point we would like to provide a bit of insight concerning the most likely role of microcomputers. Fig. 4 graphically depicts where microcomputer applications fit today, and table 1, taken from an article by Riley,[2] depicts the spectrum of computer-equipment complexity from simple hard-wired systems to high-performance general data processing equipment.

Microprocessor and microcomputer applications fall between relay logic and discrete random logic (gates and flip-flops) on one hand and small minicomputers such as the PDP 8A and the LSI 11 on the other. Microcomputers built from microprocessor chips are not as sophisticated as some of the popular minicomputers and cannot easily perform certain types of data processing problems. They are simply not set up at this time to run

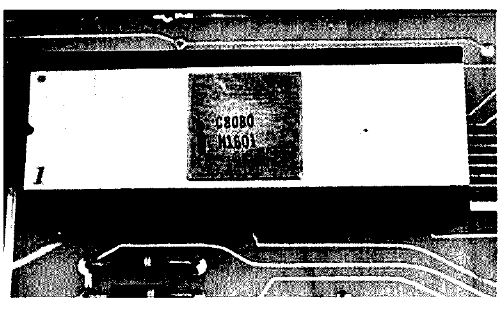


fig. 2. The 8080 microprocessor chip.

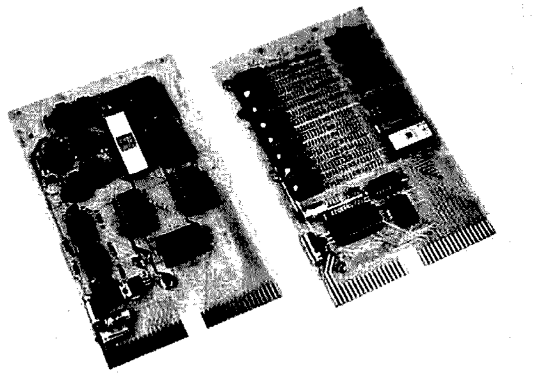


fig. 3. A typical microcomputer system. Shown on the left is the central processing unit (CPU), which consists of input/output buffer chips and miscellaneous control logic. Shown on the right is the microcomputer memory, in this case 1k of RAM and 256 words of ROM. Decoder chips permit the memory to be located anywhere within 65k of microprocessor addressable memory.

FORTRAN, COBOL, or other high-level computer languages. Those microcomputers that can, in principle, handle high-level languages still suffer in comparison with minicomputers supplied by Digital Equipment Corporation, Hewlett Packard, Data General, Varian, and other manufacturers in the amount of available high-level software.

If you want to solve tomorrow's problem, you can consider the purchase of a microcomputer system and develop your own high-level software. If you want to solve today's problem, pay particular attention to software support. Your time is valuable. If you are not careful, software costs can easily equal and exceed the total hardware costs of your data acquisition system.

For the moment, then, it would be more appropriate to call systems built around microprocessor chips *microcontrollers* or *logic processors*. They can sequence events in response to decisions upon input data. As the price of individual microprocessor chips drops from several hundred dollars per chip to $10 to $30 per chip, it will be clear that the dominant application for today's microprocessors will be as sophisticated control elements in instruments and machines of all types. We forsee minicomputer-microcomputer and computer-microcomputer hierarchies in which one to twenty instruments, machines, or devices, each containing its own microcomputer, will all be tied to a single minicomputer or computer.

the anatomy of a microcomputer

The "anatomy" of a typical microcomputer system is shown in fig. 5. This system is based upon the 40-pin 8080 microprocessor chip and possesses all of the minimum requirements for a computer:

It can input and output data.

It contains an arithmetic/logic unit (ALU), located within the 8080 chip, that performs arithmetic and logical operations.

It contains "fast" memory.

It is programmable, with the data and program instructions capable of being arranged in any sequence desired.

It is digital.

Fig. 5 shows the important data paths within the microcomputer. In the sub-sections below, we shall dissect this diagram and discuss each of the individual data paths.

Random access — A semiconductor memory into which logic 0 and logic 1 states can be written (stored) and then read out again (retrieved).

Read-only memory — A semiconductor memory from which digital data can be repeatedly read out, but cannot be written into, as is the case for random access memory.

table 1. Spectrum of computer-equipment complexity. Reprinted from *Electronics*, October 17, 1974; copyright © McGraw-Hill, Inc., 1974.

Word length	1	2	4	8	16	32	64
Complexity	hard-wired logic	programmed logic array	calculator	microprocessor	minicomputer		large computer
Application	————	control	————	dedicated computation	low-cost general data processing		high-performance general data processing
Cost	under $100	————	$1000	————	$10000		$100000 and up
Memory size	very small 0-4 words	small 2-10 words		medium 10-1000 words	large 1000-1 million words		very large, more than 1 million words
Program	read-only						reloadable
Speed constraints	real time	slow		medium			throughput-oriented
Input-output	integrated	few simple devices		some complex devices			roomful of equipment
Design	logic	logic + microprogram		microprogram macroprogram		macroprogram high-level language software system	
Manufacturing volume	large						small

Memory. Let us first consider the data communication between the 8080 central processing unit, also known as an MPU, and memory. You will require some definitions which will be useful in the ensuing discussion:[1,2]

Memory — Any device that can store logic 0 and logic 1 bits in such a manner that a single bit or group of bits can be accessed and retrieved.

Memory cell — A single storage element of memory.

Memory word — A group of bits occupying one storage location in a computer. This group is treated by the computer circuits as an entity, by the control unit as an instruction, and by the arithmetic unit as a quantity. Each bit is stored in a single memory cell.

Memory address — The storage location of a memory word.

Memory data — The memory word occupying a specific storage location in memory, or the memory words collectively located in memory.

Programmable read-only memory — A read-only memory that is field programmable by the user.

Volatile memory — In computers, any memory that can return information only as long as power is applied to the memory. The opposite of novolatile memory.

Read — To transmit data from a semiconductor memory to some other digital electronic device. This term also applies to computers and other types of memory devices.

Write — To transmit data into a semiconductor memory from some other digital electronic device. This term also applies to computers and other types of memory devices. A synonym is *store*.

The 8080 microprocessor employs 8-bit words that are stored in memory with the aid of a 16-bit memory address bus. With the aid of a quick calculation, you can conclude that there exist 2^{16} = 65,536 different memory locations which can be accessed by the microprocessor. This access to memory is direct, which means that you don't have to engage in any special tricks or

digital electronic gimmicks to access any given memory location within the 65,536 possible locations. Forty-pin integrated circuit chips do have their advantages, and this is one of them. The total memory capacity of the 8080 microprocessor is known in the trade as "64k." This is far more memory than you will ever need for most applications, but it is nice to know that you have such power in reserve.

Data is transferred between the 8080 CPU and the memory over 8-bit input and output buses, both of which are shown in fig. 5. By input we mean "input into the CPU." The term, "output," is defined in a similar fashion. Our point of reference is always the CPU. Data leaving the CPU is always considered to be "output data;" data entering the CPU is always "input data." **Fig.** 5 shows that the input and output data is transferred between the *accumulator* and memory. This is frequently the case, but in a more detailed look at the 8080 chip, you will discover that data stored in memory is transferred to other internal *registers* within the 8080 chip as well.

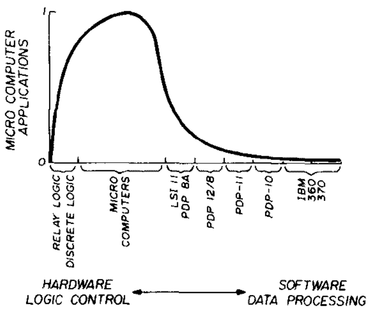


fig. 4. Where microcomputers fit: between relay and discrete logic and inexpensive minicomputers.

The most obvious such register is the *instruction register*, from which the decoding of the instruction occurs. Other registers, known as *general purpose registers* are classified by the letters B, C, D, E, H, and L. We regard the accumulator register to be the heart of the entire microcomputer. Arithmetic and logic operations are always performed to or on the eight bits of data present in the accumulator. All input and output data passes through the accumulator with the aid of two computer instructions called IN and OUT.

Between the 8080 CPU and memory there exists a single output line called memory READ/WRITE. When this line is at logic 1, you are able to READ data into the CPU either from memory or from an external device. When this line is at logic 0, you are able to WRITE data from the CPU into memory or an external output device.

As a final point, you can employ any type of "fast" digital electronic memory device, including random access memory (RAM), read-only memory (ROM), and programmable read-only memory (PROM). What do we mean by "fast" memory? Simply that the memory can perform either a read or write operation during a single microcomputer instruction. A typical 8080 microcomputer system operates at a clock rate of 2 MHz and a read or write operation takes only 3.5 microseconds. Thus, RAM, ROM, and PROM all need an access time of about one to two microseconds to allow you to take full advantage of the maximum clock speed. Slower semiconductor memories can be used, but the microcomputer will have to wait while a read or write operation takes place.

data output

The 8-bit output bus between the 8080 CPU and memory also serves as the output data bus to an external output device. When you provide output to an external device, there are several important points to remember:

> You must select the specific output device that will receive eight bits of data from the CPU.
>
> You must indicate to this device when output data is available on the output data bus.
>
> The device must capture this output data in a very short period of time, typically 1.5 μs.

The third point is perhaps the most important. Keep in mind that the microcomputer is operating at a clock rate of 2 MHz. Each computer instruction is executed in a very short period of time which ranges from 2 to 9 μs. Thus, accumulator data designated as "output data" to an external device is not available for very long. You must capture it while it is available. We will discuss the techniques that you would employ in a subsequent column; this topic is certainly among the most interesting topics that can be discussed in the field of computer interfacing.

data input

The basic considerations that apply to data output also apply to data input into the CPU from an external input device. Thus:

> You must select the specific input device that will transmit eight bits of data to the CPU.
>
> You must indicate to this device when the CPU is ready to acquire the input data.
>
> You must insure that the CPU acquires this data in a very short period of time, typically 1.5 μs.

input/output device addressing

The 16-bit memory address bus is time shared so that it can provide, at certain times, an 8-bit device identification number called a *device code*. Eight bits of information allow you to decode 2^8 = 256 different devices. When used in conjunction with two output function pulses called IN and OUT, the microcomputer system can address 256 different input devices and 256 different output devices. We might point out here that a "device" can be a complex machine such as a teleprinter or cathode-ray tube (CRT) display, or a simple device such as a single integrated-circuit chip. This is another interesting topic for discussion that we will reserve for a subsequent column.

microcomputer interrupt

Not shown in fig. 5 is a single input line to the microcomputer that generates a program *interrupt* during microcomputer operation. Such an interrupt would be generated by an external device that wishes to transfer data to or from the computer. This particular topic is quite complex, and it will be a number of months before we tackle it in this column.

The above is about the best that we can do to describe the general "anatomy" of a microcomputer in one-thousand words or less. Microcomputers are fascinating machines. They are small and relatively inexpensive, so you are less likely to be intimidated by them. They are far simpler than their minicomputer and computer counterparts and can be readily repaired by the simple process of chip substitution. They appear to be the proper answer to your childhood question, after the *Erector Set*, what?

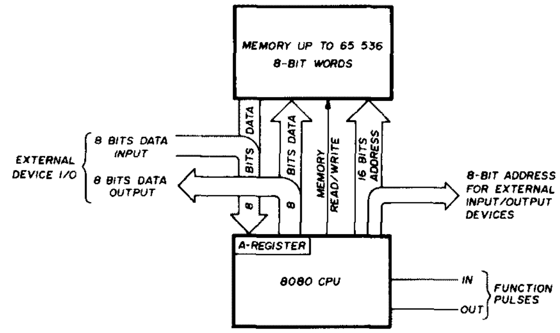


fig. 5. A typical 8080-based microcomputer system.

If you do not wish to stretch out your learning process on microcomputers for twelve months or more, we might indicate that we have just completed *Bugbook III* entitled *Microcomputer Interfacing Experiments Using the Mark 80®, an 8080 System*. It contains approximately 600 pages of text and experiments on interfacing and programming 8080-based microcomputers.[6]

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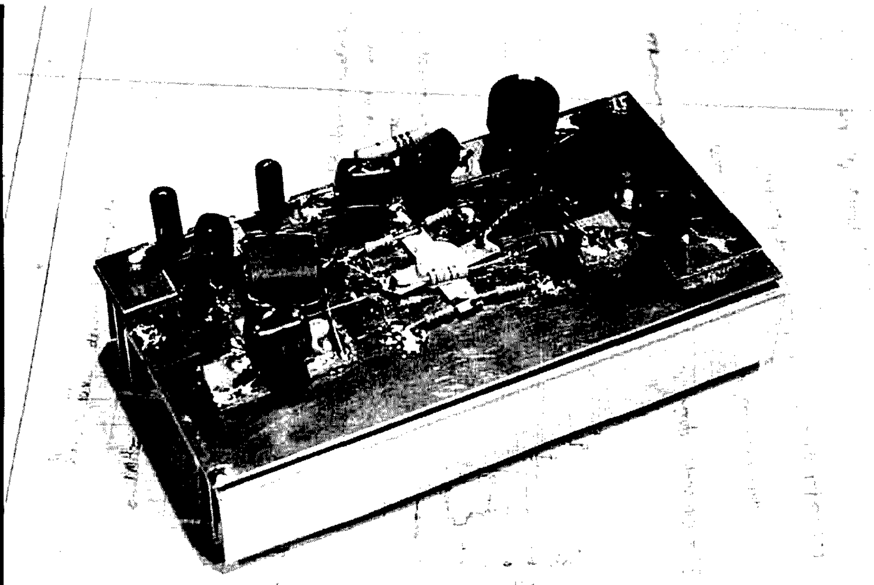
ham radio



four-watt wideband linear amplifier

A stable rf amplifier for QRP use or as a driver for higher-power linear amplifiers over the frequency range from 300 kHz to 30 MHz

There is no problem these days in building high-frequency, transistorized ssb exciters that produce outputs in the milliwatt range. However, there seems to be a dearth of information on how to get these low-level signals up to a more useful level. Articles I've seen in the amateur magazines seem mostly to use one of two extreme methods. One is to make use of the rather exotic high-priced transistors designed especially for linear power amplification; the other is to use some of the newer audio power transistors – usually with great difficulty and often with not very satisfactory results.

features

This linear amplifier uses the widely available and inexpensive (about $7.00) 2N5590 transistor to produce a power output of 4 watts across the high frequency rf range. This power level is suitable for the output of QRP rigs or as a driver for a final amplifier in the hundred-watt range.*

*This amplifier will drive the high-power linear amplifier described by Chalmers[1] to full output.

J. A. Koehler, VE5FP, 2 Sullivan Street, Saskatchewan, Canada S7H-3G8

The amplifier gain is flat over the high-frequency rf range, being only 3 dB down at 300 kHz and 30 MHz. In fact, the amplifier still produces useful gain at six meters. The exact gain will depend somewhat on the transistor used, but the version I built had a midband gain of 22 dB. This means that full output on the amateur bands up to 15 meters can be obtained with only 25 milliwatts of drive; 40 milliwatts is required at ten meters. The amplifier output may be either shorted or left open indefinitely with no damage even with full drive. The amplifier is also very stable and shows no tendency to oscillate.

circuit

The amplifier schematic is shown in **fig. 1**. The stability and wide frequency response are achieved by adding considerable negative feedback to an otherwise conventional broadband amplifier. The small inductance in series with the 560-ohm feedback resistance decreases the feedback at the higher frequencies. The exact value is not critical. About 25 turns of number 30 (0.25mm) wire closewound on a ½-watt resistor will do very well.

The amplifier operates in the class-A mode, and the transistor has a quiescent power dissipation of 5 watts. A fairly efficient heat sink is required. While commercial heatsinks are good, an acceptable one can be made from three sheets of 0.06-inch (1.5mm) aluminum formed and assembled as shown in fig. 2. After all holes are drilled, they should be deburred so the pieces will make good contact with each other. It's a good idea to put silicone grease or heatsink compound between the pieces before final assembly.

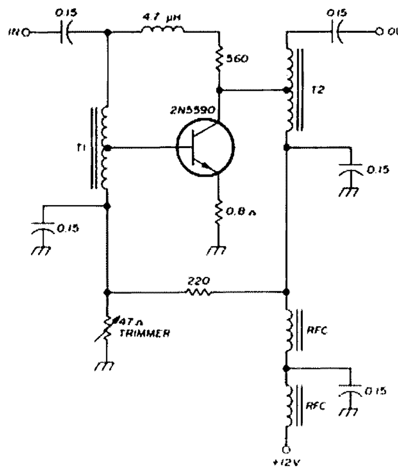
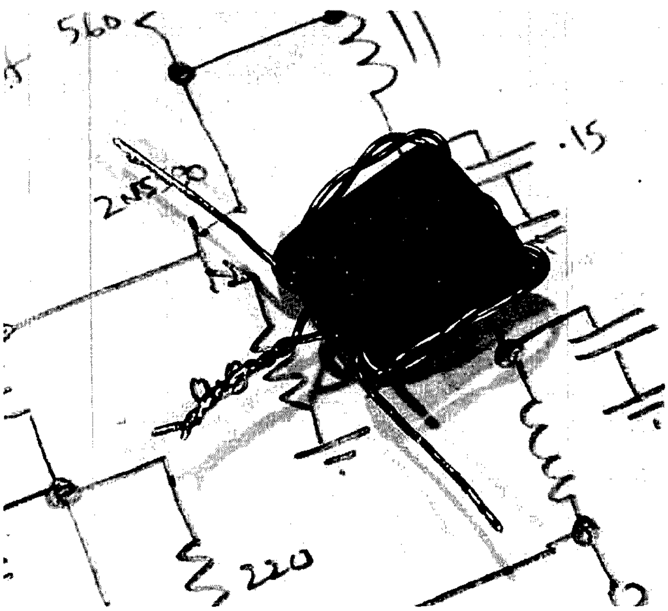


fig. 1. Rf amplifier schematic. T1, T2 are wound on two-hole balun cores as found in TV-set input circuits. Emitter resistor is made from four 3.3-ohm, ¼-watt resistors in parallel. All capacitors are 100-volt plastic; rf chokes are 1 or 2 turns through a ferrite bead.

One of the broadband transformers used in the input and output circuits.



construction

The circuit is built on a piece of single-sided circuit-board material mounted copper side up on the heatsink. A large clearance hole for the transistor is drilled in the center so that the transistor can be mounted directly on the heatsink. It's important to put silicone grease or heatsink compound on the mounting surface of the transistor before assembly. It's also important *not* to overtighten the transistor mounting nut. Components are mounted between pads of PC material approximately 0.39 x 0.39 inch (1 x 1cm) cemented to the main circuit board. Pad locations may be found by laying out the components you wish to use on the board. The general appearance of the amplifier (before adding components) is shown in fig. 3.

The transistor tabs are fragile, so the transistor should be mounted in its final position first and the components soldered to the tabs later. *Do not* reverse this order or the tabs will be stressed when you tighten the tran-

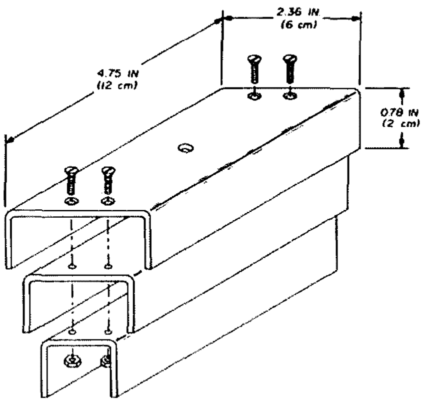


fig. 2. Method of constructing transistor heatsink. Aluminum sheet is formed to approximate dimensions shown, then assembled. Hardware is 4-40 (M3). Top holes are countersunk. Transistor stud clearance is drilled after assembly.

sistor mounting nut. The 0.8-ohm emitter resistor must have a very low inductance, which may be achieved by paralleling several higher-value resistors. I used four 3.3-ohm, ¼-watt resistors soldered symmetrically between the emitter tabs and the ground plane as shown in the photograph.

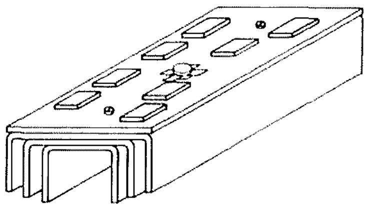


Fig. 3. Printed-circuit board and heatsink mounting details. Mounting pads for components are made of PC board material.

The input and output transformers are wound on two-hole balun cores as found in TV sets. The ones I used are manufactured by Phillips; their type number is 4322-020-31520. The windings are made by twisting two pieces of number 22 (0.6mm) enamelled wire together about three twists per inch (one twist per cm). This twisted pair is then wound through the core as

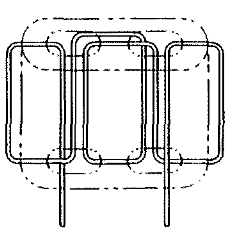
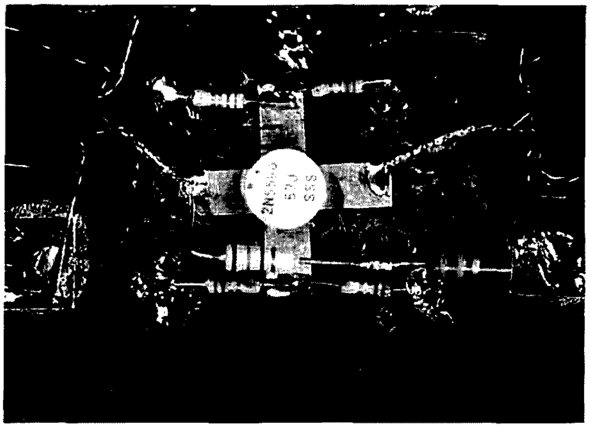


fig. 4. Transformer winding details. No. 22 (0.6mm) enamelled wire twisted pair is wound through a two-hole balun core.

shown in fig. 4 and the photograph. Using an ohmmeter to identify the wires, one end of one wire is connected to the opposite end of the other wire, which is the center tap of the transformer. The wide frequency response of the amplifier is due to these transformers, and the general method of construction should be followed, although wire size, number of twists and number of turns through the core are not too critical.

Chokes in the main supply line are made by winding one or two turns through any of the widely available ferrite beads. I used one turn through a two-hole bead for each of the chokes. Large values of inductances are not required here since the power-supply line operates at very low impedance.



Underside of amplifier. Emitter-resistor assembly is shown paralleled between emitter tabs and ground plane. Input and output transformers are at extreme right and left.

No tuning is required, so the amplifier is made ready for use merely by adjusting the quiescent current level. First set the 47-ohm trimmer to minimum resistance then connect the 12-volt power supply and adjust the trimmer so that the total amplifier current drain is 0.4 ampere.

At the one-watt output level, the second harmonic measured 30 dB down with respect to the fundamental. I wasn't able to measure the intermodulation distortion with the test equipment I had available. From the measured performance of similar amplifiers,[2] I'd expect it to be about -40 dB. For CW use, the quiescent current may be lowered to reduce wasted power; however, the output harmonic content will increase and the overall gain will decrease.

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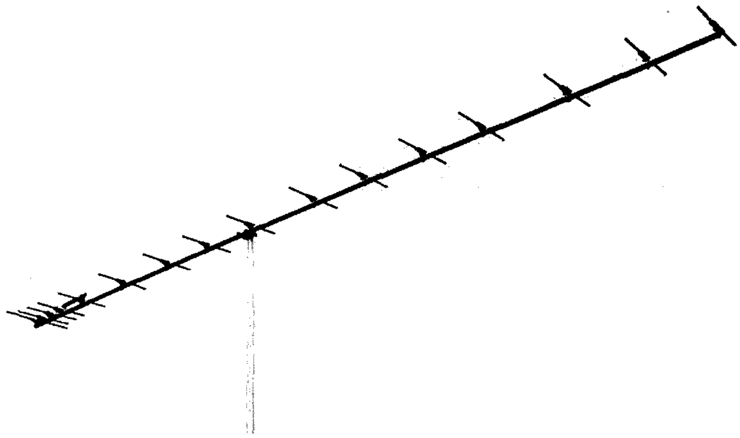


high gain yagi

for 432 Mhz

A new long-boom 16-element Yagi design for 432 MHz that provides 15 dB gain over a dipole

For years the amateur uhf community has been trying to come up with a reproducible, high-gain Yagi beam for 432 MHz. At one time it was generally agreed among amateurs that the dimensions of a really long uhf Yagi antenna were so critical that it was impossible to build a practical, reproducible, high-gain, multi-element beam, and most uhf operators switched to the less critical co-linear array. Unlike the long-boom Yagi, the colinear is a low-Q antenna, so *none of the dimensions are overly* critical and it is easily reproduced for uhf operation.

However, as has been pointed out by Ed Tilton, W1HDQ,[1] it is possible to build Yagi antennas for 432 MHz (and other uhf frequencies) if *all* dimensions are properly scaled. Most experimenters scaled element length and spacing, but failed to scale either the element or boom diameter – this resulted in antennas that exhibited little more gain than a dipole, or worse. W1HDQ's 11-element, 432-MHz Yagi design was the first that proved to be reproducible, and although it uses a wooden boom, large numbers are being used by amateurs on the 432-MHz band. The gain of the Tilton Yagi has consistently measured about 13 dBd (gain over a dipole).

Other successful 432-MHz Yagi designs are those of WØEYE[2] and K2RIW.[3] WØEYE's 15-element design, which uses a 10-foot (2.9m) metal boom, attracted wide attention, but not everybody who tried to build it was successful. K2RIW's 13-element Yagi, which uses insulated elements (8-foot [2.4m] boom), has been quite popular in the East, and has consistently been shown to provide about 15 dBd gain.

Described here is another long-boom Yagi for 432 MHz which provides about 15 dBd gain. This has been confirmed at antenna measurement contests on both the East and West Coast. This Yagi, which was designed by Mike Staal, W6MYC, and Mel Farrer, K6KBE, of KLM Electronics, is based on successful design techniques

By Ken Holladay, K6HCP, 2140 Jeanie Lane, Gilroy, California 95020

proven on hf and vhf and uses a broadband driven structure which consists of three elements (**fig. 1**). This provides a reasonable operating bandwidth and ease in coupling to the 12 directors and one reflector. The broadband structure, in addition to providing optimum coupling to the directors, is the key to reproducibility. Small variations in dimensions can be tolerated without significantly changing the operating characteristics of the antenna.

construction

As is shown in **fig. 1**, the antenna is based on a 1-inch (25cm) diameter boom, 12-feet (3.7m) long. Each of the elements is 3/8-inch (9.5mm) diameter aluminum tubing, insulated from the boom except for the single mounting screw (this type of element mounting *must* be used for the dimensions given in fig. 1). The driven elements are cross connected using ¼-inch (6.5mm) wide aluminum strap. The feedpoint impedance is 50 ohms (balanced) and must be connected to a balun using low-inductance copper strap 5/16 inch (8mm) wide.* To prevent aluminum-to-aluminum and aluminum-to-copper corrosion, all joints should be coated with *Penetrox A* or equivalent weatherproofing. An acceptable balun can be made as described by K6HCP and WA6GYD in the ARRL *Radio Amateur's VHF Manual*.[4]

performance

At my station I have two of these antennas mounted side by side, and they have provided the expected results. Los Angeles is about 300 miles (483km) away, over mountainous terrain, and good solid contacts on 432 MHz are a nightly occurrence. Activity on 432 is starting to increase, and I feel confident that this new antenna, which is easy to build, will do a great deal to stimulate growth on this band.

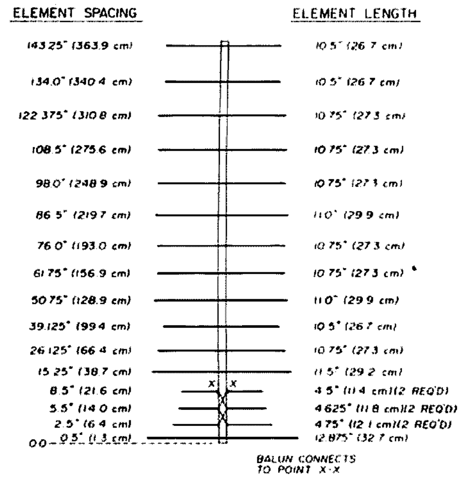


fig. 1. Layout of the 16-element Yagi for 432 MHz. Elements are 3/8" (9.5mm) diameter aluminum tubing, insulated from the boom except for the single mounting screw as shown in fig. 2.

*For those readers who do not have the time or material to build their own, this antenna is available from KLM Electronics, 17025 Laurel Road, Morgan Hill, California 95037.

references

1. Ed Tilton, W1HDQ, "Antennas for 220 and 420 MHz," *The Radio Amateur's VHF Manual*, ARRL, Newington, Connecticut, 1972, page 208.
2. Don Hilliard, WØEYE, "15-Element Yagi for 432 MHz," *QST*, January, 1972, page 96.
3. Dick Knadle, K2RIW, "13-Element Insulated Yagi for 432 MHz," *The ARRL Antenna Handbook*, 13th edition, 1974, page 243.
4. Ken Holladay, K6HCP, and Don Farwell, WA6GYD, "Making and Using Baluns," ARRL *Radio Amateur's VHF Manual*, page 170.

ham radio

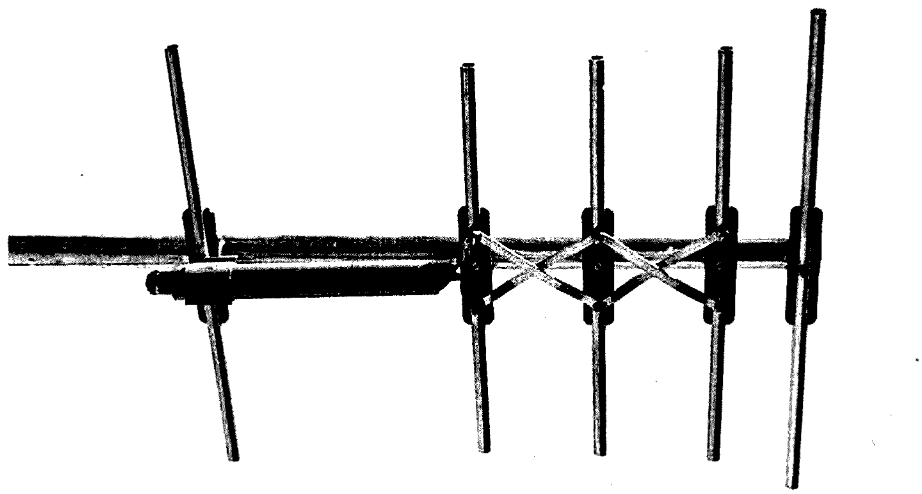
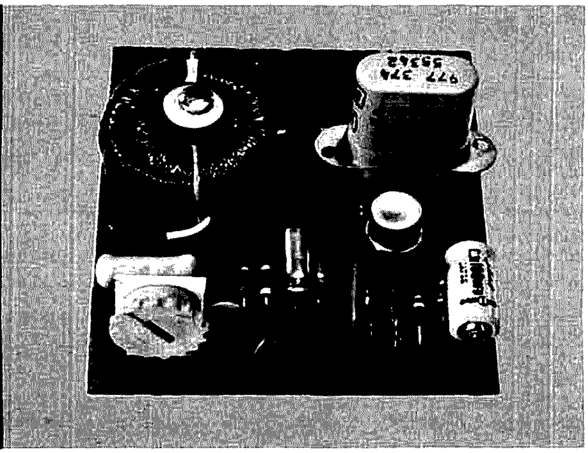


fig. 2. First five elements of the high-gain 432-MHz Yagi, showing element mounting and balun installation. The three-element, cross-connected driven structure is at the center.



telephone controller

for remote repeater operation

Using a modified RTTY autostart circuit to activate a repeater without a telephone

An earlier issue of *ham radio* described a telephone controller that could be used to turn any device on and off, such as a repeater.[1] Hopefully you've already read that article, so we won't go into the actual telephone controller; instead we'll show you how to operate the controller without a telephone. Many have asked us how this can be done, so we designed a circuit that will decode a tone and change it to a dc voltage that will allow the controller to be operated without a telephone.

We already had a circuit board that would do this.* This circuit was originally designed for RTTY *autostart*, and by simply retuning it to a chosen audio frequency, we found that it worked perfectly as a tone decoder. The RTTY autostart circuit is shown in fig. 1. All we had to do was omit the relay, bring out the lead from Q2 collector, and route it to the telephone controller junction of U1, pins 14 and R8.[1] On the original telephone controller you can omit Q1, Q2 and Q3; CR1, CR2 and CR3; R1-R7; C1-C4; and K1. (These parts were used to validate the telephone line only.)

*Available from Circuit Board Specialists, 3011 Norwich Avenue, Pueblo, Colorado 81008. RTTY autostart printed circuit board $3.50 each. Complete kit, less power supply (+18 to 24 Vdc) $14.50. (Specify approximate tone frequency.)

The circuit is a tone decoder that turns transistor Q1 on and off. A tone of your choice is fed into the circuit from your receiver through R1 and decoupling capacitor C1. The tone is decoded by L1, C2. Q1 is a voltage amplifier. The amplified tones are rectified by diodes CR1 and CR2. The resulting dc voltage is fed to the base of Q2. Capacitor C7 requires about two seconds to charge and discharge, resulting in Q2 turning on and off at a rate similar to the ring rate from the telephone, as decoded by the original circuit in the telephone controller. Therefore, assuming the telephone controller circuit has been properly programmed, one would ring three times, hang up, wait twenty seconds, and ring three more times. The same thing would be done with the tone encoder on your mobile or base rig; i.e., push the button three times for one second each (or longer); stop; wait twenty seconds; then push the tone button three more times.

The only thing that will take a little time is tuning the toroid for the audio frequency you desire, which is done with the aid of an oscilloscope or vtvm on the ac volts scale.

Put the plus lead of the scope or vtvm to the gate of Q1. Dc voltage need not be applied to the circuit. Apply your desired tone to the input and open R1 all the way. Adjust C2 for maximum ac volts or peak-to-peak voltage on the oscilloscope. Remember, the better the tuning, the narrower will be the bandpass.

After it's all hooked together, apply a dc voltage between +12 and +24 volts to the decoder board and

By Robert C. Heptig, KØPHF, Robert D. Shriner, WAØUZO

Robert C. Heptig, KØPHF, P.O. Box 969, Pueblo, Colorado 81002, Robert D. Shriner, WAØUZO, 1740 E. 15th St., Pueblo, Colorado 81001.

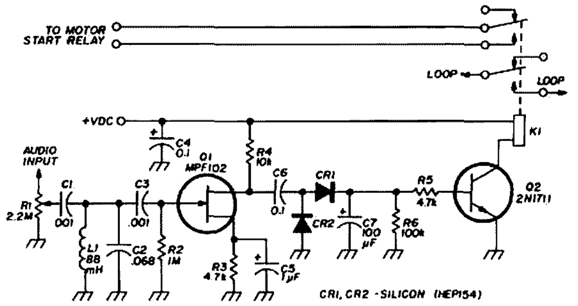


fig. 1. RTTY autostart circuit, which is used with IC U1, the SN7490 in the original article (reference 1), as a tone decoder. A tone of your choice is fed into the circuit through R1, C1 and is decoded through L1, C2. K1 is a Potter & Brumfield SC-4332.

telephone controller board, then test it as described in reference 1 (with the exception of the validating circuit). Adjust R1 of the decoder board to allow just enough audio to do the job with respect to the amount of deviation of your tones.

tone access

Just for the fun of it we added a tone access for your repeater to the circuit. Simply put the relay back into the circuit of the RTTY autostart as shown in fig. 1 and wire it into your repeater as shown in fig. 2. In this configuration the COR cannot be keyed unless a brief tone is applied to your carrier, which will cause the modified RTTY autostart circuit to provide a ground for

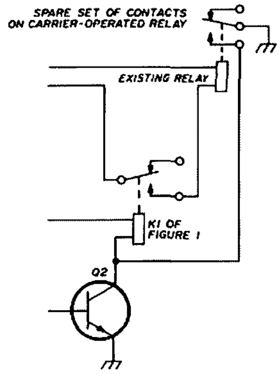


fig. 2. Tone-access circuit for repeaters. A brief tone applied to your carrier causes the modified autostart circuit to ground your COR as long as the carrier is present.

your COR; then a spare set of contacts on your COR will maintain ground as long as your carrier is present. By tuning the decoder as a broadband amplifier (increase values of C1 and C3), this circuit could be used for a vox-operated repeater. With a little imagination, you can probably come up with many possibilities for this decoder.

reference

1. Robert C. Heptig, KØPHF and Robert D. Shriner, WAØUZO, "Automatic Telephone Controller for Your Repeater," *ham radio*, November, 1974, page 44.

ham radio

repair bench

Michael James

basic troubleshooting

Troubleshooting and equipment repair are two of the toughest problems faced by radio amateurs today. Part of the difficulty is due to the fact that modern ssb equipment is much more complex than the old a-m and CW gear of twenty years ago, but perhaps more important, few amateurs build major pieces of their station equipment anymore so they are probably not as familiar with its circuitry as they should be. When your receiver or transmitter starts giving you trouble, more than likely it will be returned to the factory to be repaired. If the problem isn't too severe, you may avoid using that function which is affected or overlook it altogether. In some cases you may not even be aware of a problem unless another amateur brings it to your attention (distorted speech, poor sideband suppression or splatter, for example).

Although there *may* be some equipment repair problems that are best sorted out by the factory, in most cases you can save yourself a lot of time and money by fixing it yourself. Once you send your gear back to the factory, you may have to wait a month or more until you can get back on the air. In addition, you will probably have to pay the factory ten dollars an hour or more for their technician's time.

Troubleshooting electronic equipment is not difficult, nor does it require a bench full of test equipment. A large selection of test equipment may simplify the task, or allow you to solve a problem more quickly, but 90 per cent of all troubleshooting can be accomplished with a volt-ohmmeter and other simple test equipment you already have on your workbench. In those cases where you need a calibrated signal generator or an oscilloscope, you can often borrow one from your local radio club or from an amateur who lives nearby.

In the coming months this column will be devoted to troubleshooting techniques and how you can use them to fix your own equipment. Although much of the initial discussion will be in general terms that are applicable to practically any electronic equipment, future columns will discuss specific pieces of equipment and unique or unusual circuitry that requires a somewhat different procedure. If you have solved a particularly difficult equipment problem, we would like to hear about it. There may be others who will be helped by your success.

basic troubleshooting

There are three basic troubleshooting techniques which can be used to locate and fix circuit malfunctions: signal tracing, resistance measurements and voltage measurements. In receivers and transmitters the problem area is usually located with signal tracing, then pinpointed with resistance and/or voltage measurements. Although some electronic circuits such as gain-control circuits don't lend themselves to signal tracing, the majority of receiver and transmitter circuits can be quickly checked with this method. Once you know how to use signal tracing, in fact, you will probably agree that it's one of the quickest ways to track down a circuit problem.

Basically, signal tracing consists of injecting a signal at the input to a piece of equipment and checking its path through the equipment. If the signal appears at the input to a stage, but not at the output, that stage is the culprit. It may not be the only culprit, but once it's been fixed, you can locate other problem areas along the signal path.

The signal tracer is essentially a very quiet, high-gain audio amplifier with headphone or speaker output. One commercial version which is available at modest cost is shown in the accompanying photograph. If you wish, you can build a simple high-gain audio amplifier around an op amp IC as shown in **fig. 1**, and in a pinch you could even use one channel of your

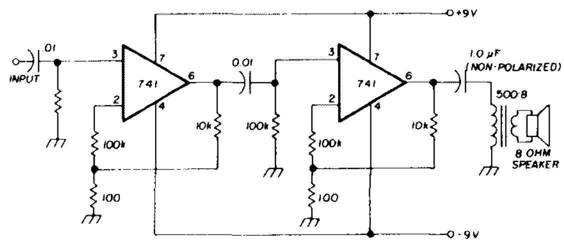


fig. 1. Signal tracer circuit which is based on the 741 op amp ICs. Gain is about 80 dB; audio power output is approximately 40 mW, sufficient for most signal tracing.

stereo system. This is all you need if you're working with audio systems, but if you're troubleshooting rf and i-f stages, you will also need a simple demodulator probe such as that shown in **fig. 2**. The one I use is built into a discarded plastic ballpoint pen. You can also use one of the rf probes which are available for vacuum-tube voltmeters.

In addition to the signal tracer (audio amplifier and rf probe) you will also need a signal injector – a device which has a broadband signal output from audio through vhf. There are several pencil-sized signal injectors on the market for less than ten dollars. Most consist of a simple 1 kHz multivibrator which has high harmonic content well above 30 MHz. The circuit in **fig. 3**, which uses

inexpensive high-speed switching transistors, can be used for signal tracing through at least 50 MHz. Built on perfboard, this unit is small enough to fit inside the aluminum cases in which expensive cigars are sold (you could also use a plastic pencil holder or toothbrush case).

signal tracing

Whatever kind of signal tracer you decide to use, you'll want to get the most out of it. Many people who already use signal tracers seem to think that signal tracing is limited to localizing trouble in one section of a receiver or transmitter. However, as will be shown later,

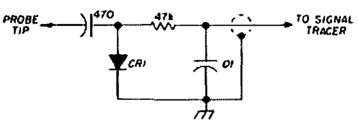


fig. 2. Demodulator rf and i-f probe for signal tracing. Diode CR1 can be practically any signal diode such as 1N34A, 1N60, 1N67A, etc.

the signal tracer can also be used to pin down defective components. All you have to do is know how to use it.

Fast troubleshooting with a signal tracer demands logic, and you'll have to supply that. But even if you haven't done any troubleshooting before, you'll be *amazed* at how quickly you can track down a faulty circuit with a signal tracer. Fixing the bad circuit after you've located it may be another story, but if you use logic, and the resistance and voltage measurements we will discuss in future columns, you can probably repair any electronic circuit ever built. Things are simplified tremendously if you have a copy of the schematic or the manufacturer's maintenance instructions, but even without these you can, with persistence, be successful.*

As a starter I'll show you how to use a signal tracer to troubleshoot the sophisticated amateur communications receiver shown in **fig. 4**. This block diagram is fairly typical of modern superheterodyne receivers although some models may use only one frequency-conversion stage, or may not be equipped with a crystal calibrator or a separate a-m detector. I should also point out that it doesn't make any difference if your receiver uses vacuum tubes, transistors, or some combination of these – the basic signal tracing technique is the same.

First of all, take a look at the schematic and mentally divide it up into blocks representing each stage or function in the set. **Fig. 4** has been divided into four basic sections: rf, high i-f, low i-f and audio. In some cases you might want to consider the detectors separately, but they are usually included as part of the last i-f.

First, the rf section. When signal tracing here you'll have to use the demodulator probe. If the receiver is connected to an antenna you will hear a mismash of incoming signals because most receivers don't have sufficient front-end selectivity to pick out any one signal –

*Manuals for most amateur equipment manufactured between 1940 and 1965 are available from Hobby Industry, W0JJK, Box H864, Council Bluffs, Iowa 51501. Send self-addressed, stamped envelope for quote.

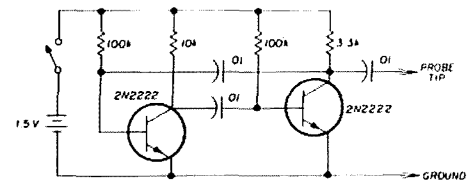
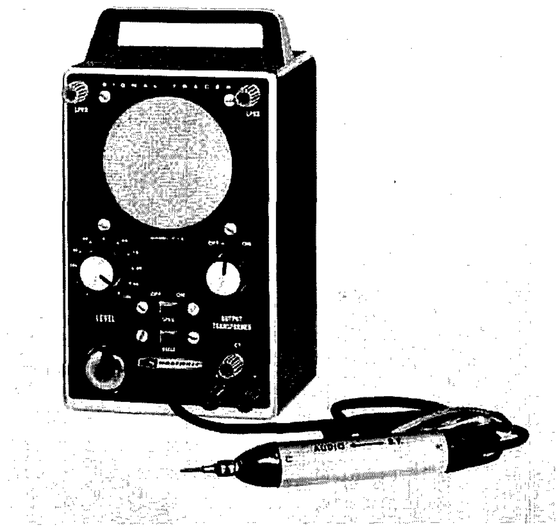


fig. 3. Signal injector is basically 1 kHz multivibrator which has high harmonic output. Circuit shown here has usable output up through 50 MHz. Practically any npn transistors will work in this circuit.

that's done further on, in the low i-f. If all the rf stages are normal, once you set the bandswitch all the signals within several-hundred kHz will be heard through the signal tracer. The collectors of the rf amplifier and mixer transistors (plate circuits in vacuum tube receivers) are the points to check with your probe. If you don't get any signal output from the mixer, something in the rf section is dead.

The high i-f processes the output of the first mixer and consists of a bandpass filter, the second mixer and the variable frequency oscillator. If any of the circuits in the high i-f isn't working properly, the signal picked up by your tracer at the output of the second mixer will reflect it. The low i-f includes the selective filter, i-f amplifier amplification and the detectors. You'll need your demodulator probe for the i-f stages, but the quickest test point for the whole section is after either of the detectors. Here you should hear a clear, undistorted audio signal without the probe. The audio section can also be checked without the probe. If the receiver is okay, you should hear a nice strong signal at the output of the last audio stage.

If the receiver isn't working properly, the quickest



Heathkit IT-12 signal tracer has both visual (eye tube) and audio output. A switchable audio-rf probe is included with the kit, which sells for $32.95.

way to find the bad circuit is to check signal output about halfway through the set. A good point is the output of the second mixer. If the receiver is connected to an antenna the signal you hear should change as you tune the vfo (since the demodulator probe is an a-m detector, ssb signals will be unintelligible). If your receiver will tune to one of the WWV channels, this makes an excellent test signal, or you can use your signal injector. The pitch of the wideband injector signal, however, will not change as you tune the vfo.

If the signal is okay at the output of the second mixer, you have cleared the front-end circuits of any suspicion and can proceed to the last half of the set – the output of the low i-f amplifier is a good point. If you don't get an output from the second mixer, the low i-f and audio sections are probably okay.

Assume you get nothing at the output of the second mixer. Divide the front end roughly in half and use the tracer and demodulator again. The output of the first mixer is a good check point. If you have the proper signal there, there's something amiss in the bandpass filter, vfo or second mixer. If there's no signal output from the first mixer, the rf amplifier, crystal oscillator or first mixer stage must be at fault.

The last half of the receiver can be attacked with similar logic. If the signal was okay at the second mixer, the next logical dividing point is the output of the detector, which can be checked directly, without the probe. A signal in the tracer means that everything is okay up to there and the trouble is in the audio section. If you don't get a signal, check the output of the other detector. No signal means it has been blocked between the second mixer and the detector – the crystal filter or one of the i-f amplifiers is the problem.

Note that with only two signal tracer checks you have ioslated the problem to one small, functional section of the receiver. If the signal is okay at the input to a stage and not at the output, it's obvious the trouble is between those two points. It's a simple matter to check each of the individual stages within a section to pinpoint the offending one.

The divide-and-conquer technique of stage isolation works just as well for other symptoms as it does for a radio that is completely dead. You can hunt noise or hum, for example, tracking down the stage where the trouble first appears. It also works for distortion.

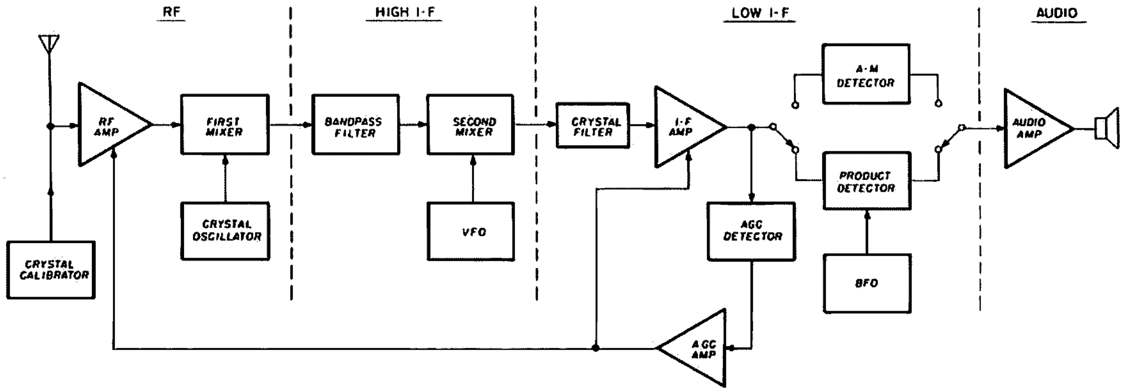


fig. 4. Sectionalizing an amateur communications receiver by functions. Dividing it up this way makes it easy to track down trouble with the signal tracer.

other checks

If the receiver is suffering from poor sensitivity, the problem can be signal traced by the "straight through" method. If reception is poor, the fastest way to determine which amplifier isn't doing its job is to check the gain of each stage by touching the signal-tracer probe to the input and output; if there is little or no increase in signal strength, the amplifier is weak. Although transistor mixer stages usually have some gain, vacuum tube mixers seldom exhibit gain and may often have a small signal loss, so keep this in mind. The filters introduce loss, too, but you can judge if it's too much after you have a little practice.

There are other little tricks of troubleshooting logic that make it easy to find troubles. If your receiver works alright on a-m but not on ssb or CW, for example, the difficulty is probably with the product detector or bfo –

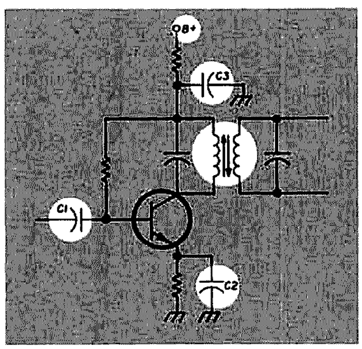


fig. 5. You can check these components with your signal tracer without even unsoldering them from the circuit.

they are the only stages which are not common to a-m. If weak signals sound okay, but strong ones distort, a good suspect is the agc stage which may not be controlling the rf and i-f gain as it should, letting strong signals overload the receiver. Likewise, frequency jumping or drift can usually be traced to the vfo; audio distortion eliminates all but the detector and audio stages; and poor selectivity is usually caused by a bad crystal or mechanical filter.

getting closer

After they've pinpointed the stage which is causing the problem, many technicians put away their signal tracer and reach for their voltmeter. However, the signal tracer can still tell you things about the circuit you can't find out with a voltmeter. In the amplifier circuit of **fig. 5**, for example, the highlighted coupling and bypass components can be tested right in the circuit without even unsoldering them.

The coupling capacitor, C1, and the interstage transformer, T1, should pass the signal along with very little attenuation. Whether they are large, as in audio stages, or small, between rf or i-f amplifiers, there should be about the same amount of signal on both sides. If there is any attenuation, it should be small. To check, touch the tracer probe to the input side of the component, then to the output side – if the output is much weaker than the input, the part is defective.

The bypass capacitors, C2 and C3, shunt the signal to ground and their values are chosen to short out practically all the signal at the emitter (C2) and at the power supply end of the interstage transformer (C3). The tracer should hear very little signal at either point. If there's any substantial signal the capacitor isn't doing its job. Even if the transistor is in good health, bad bypass capacitors at C2 or C3 will seriously degrade the gain of the stage.

Sometimes, when checking stage gain or components, you'll find that you don't have sufficient signal strength to determine if a component is doing the job it should. In this case it's helpful to place the signal injection directly at the input to the stage. This will bring the signal level up to the point where you can make meaningful measurements. You can also use the signal injector to quickly move through the receiver to determine which stage is causing the problem. Simply touch the probe of the signal injector to the input of each stage, starting at the audio output stage, and move back toward the front end, stage by stage. If everything is working properly you will hear the 1 kHz modulation through your receiver's speaker as you inject signal into each stage.

Finally, you can check the B+ line with your signal tracer for any traces of hum. Power supply filter capacitors are like any other bypass capacitors in that they should shunt all signal voltages to ground (power supply ripple in this case) and leave only pure dc. If one of the filter capacitors is weak, you'll hear a considerable amount of hum in the signal tracer. If the dc line isn't properly decoupled you may hear a whistling or hissing sound that is an rf or i-f signal if you could unscramble it. This can usually be traced to a bad bypass (decoupling) capacitor somewhere along the B+ line.

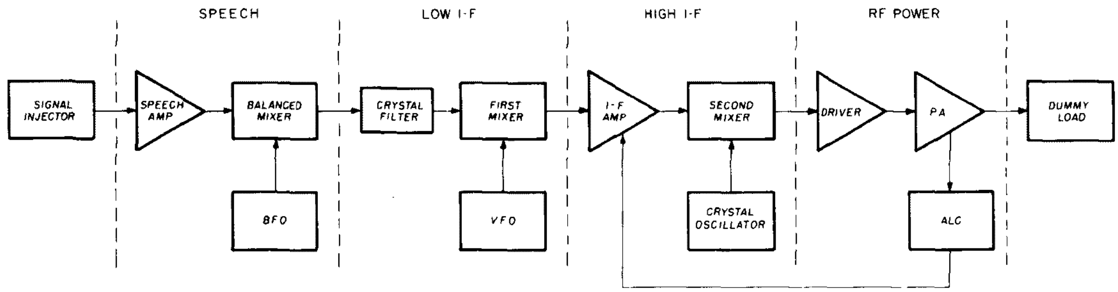


fig. 6. Modern ssb transmitter can be sectionalized by function for troubleshooting purposes.

transmitter signal tracing

A modified form of signal tracing is also suitable for tracking down problems in ssb (and a-m) transmitters. In this case the signal injector is connected to the microphone jack and the transmitter is terminated in a dummy load as shown in **fig. 6**. Except that the position of the stages is reversed (audio front-end, rf output), the functions of the various stages in a modern ssb transmitter are not that much different than those in a superheterodyne receiver.

By using the signal tracer to track through the stages of the transmitter, you can quickly locate a stage which is blocking the signal (use the demodulator probe for the balanced modulator output and all following stages). The rf output from the final amplifier may be a little too much for the detector diode in the probe so don't connect it directly to the *output* – placing the probe tip next to the power amplifier compartment should provide enough signal for tracing purposes.

Although the signal tracer won't track down distortion, poor sideband suppression, or vhf parasitics in the transmitter, it's useful for quickly isolating a nonfunctioning stage or component. The signal tracer can also be used to eliminate hum and locate bad decoupling capacitors which are causing unwanted rf feedback. Other transmitter troubleshooting techniques will be discussed in a future column.

ham radio

RAM keyer update

Circuit improvements for the random access memory electronic keyer described in a previous issue

This is a followup report on the two-RAM programmable keyer article in the October, 1973, issue of *ham radio*[1] and my correction note[2] in the December, 1974, issue. Many inquiries have been received concerning possible parts procurement and solutions for faulty keying. This article will enable you to build this keyer with a minimum of frustration.

printed-circuit board

The majority of inquiries were about the procurement of the printed-circuit board. As indicated in the original article, the board, as well as the kit of parts for the keyer, could be obtained from the indicated address. It now appears that this company is no longer the source for the parts or the printed-circuit board. If you have a lot of time and patience the circuit can be hand wired. In fig. 4 a full sized view of the etched board, from the foil side, is shown for those who would like to build one. The layout of the components and external connections are shown in fig. 5. This board diagram is free from the errors of the original version and incorporates the circuit change as described in reference 2.

clock circuit

Clock pulses are derived from a NE555 timer connected for astable operation; i.e., as a free-running multivibrator. Using the values of the original article, the maximum theoretical keyer speed will be about 17 wpm. Redesign of the keyer for higher speeds is easily accomplished by noting fig. 1. The IC manufacturer gives the clock frequency in terms of R_A, R_B, and C as:

$$f(Hz) = \frac{1.44}{(R_A + 2R_B)C}$$

where R_A = 1k ohm, R_B = 6.8k to 56.8k ohms and C = 6.8 μF. Converting the clock frequency to keying speed,[3]

$$Speed\,(wpm) = 1.2f$$

$$Speed\,(wpm) = \frac{1.73}{(R_A + 2R_B)C}$$

so that the speed range is expected to be from 2.2 to 17.4 wpm. Fig. 2 shows the values of C and $(R_A + 2R_B)$ for the desired speeds. Using C = 6.8 μF the graph shows that the speed varies when the resistance changes from 14.6k to 114.6k ohms. In my case, a 3.3k resistor was used in place of the 6.8k, and a 5 μF capacitor was used. Maximum keying speed was then 32 wpm, and a slight reduction in the duty cycle of the clock pulses (8%) occurred, which didn't affect the keyer performance. The 50k pot should have a log rather than a linear taper to permit a linear speed range; otherwise the higher keying speeds will crowd together near the upper portion of the pot rotation.

random-access memories

The second largest number of inquiries was about the RAM devices. The (Signetics) 25L01B is the low-power dissipation, pin-for-pin equivalent of the popular 1101 256-bit RAM (National Semiconductor and others). I first used the 25L01B* without any problems. If its price is a little too high for you, you can try the 1101 version as advertised by large discount houses in the amateur literature (about $2.50 each). My experience has been that you get what you pay for. I bought a half dozen of these bargain specials and only one worked correctly. If you expect a cheap bargain, you'll probably get a cheap device. *Caveat emptor*.

faulty keying

Even after incorporating the changes in the correction note,[2] some readers still had problems with sending code

*Obtained from Schweber Electronics, 5640 Fisher Lane, Rockville, Maryland 20852 at $6.50 each.

By Howard M. Berlin, K3NEZ, 2 Colony Boulevard, Apt. 123, Wilmington, Delaware 19802

characters. This annoying problem arises from stray rf and spikes generated from the TTL logic. In his article on the Accu-Keyer,[4] WB4VVF discusses some possible cures. *It is essential that all external leads be shielded* from rf. Use RG-174/U or similar coax from the keyer output to the transmitter. If an external paddle is used, use shielded three-wire cable from the paddle to the keyer. As a further precaution, add 0.1 μF bypass capacitors on the three inputs of the paddle at the input jack. TTL spikes can usually be eliminated by adding 0.01 μF capacitors from each IC chip's +5 volt pin to ground. In more stubborn cases it may be necessary to place a number of 0.01 to 0.1 μF capacitors around the edge of the printed circuit board (ground) to +5 volt points. I used about eight additional capacitors and have the keyer right next to my kilowatt linear without any trouble in keying.

Another tip on bypassing to cure faulty keying was received from Ken Beck, K3DW. He found that false

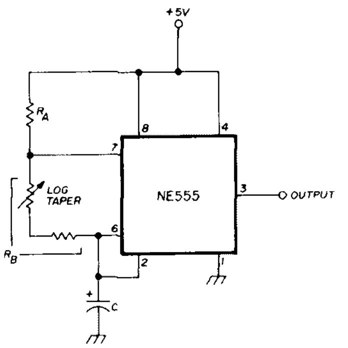


fig. 1. NE555 timer IC connected for astable operation.

dash generation occurred due to transient triggering of the master flip-flop in the 7473 IC. To eliminate the transients requires the addition of a disc capacitor bypass (0.02 to 0.05 μF) directly between pins 4 and 11 of this 7473. Similar bypassing of the 7473 that controls the address cycle also helps to prevent unwanted cycle starts caused by transients. Also disc capacitors (0.02 μF) connected from each key lead to chassis (installed right at the key jack) helped to reduce false triggering. In any case, bypassing is necessary to eliminate keying transients.

Another possible cause of faulty keying is in the clock circuit. As mentioned before, the clock is free-running and will continue to run until power is disconnected. Faulty keying *may* occur if the pulses of the individual Morse code characters are not in synchronization with the rest of the logic. The only way to cure this is to redesign the clock to run only when the desired characters are being sent.

momentary clear switch

A useful addition to this keyer circuit, offered by K3DW, is a momentary switch to clear the memory during either read or write operation, **fig. 3**. The 6.8-μF timing capacitor for the NE555 clock IC is grounded

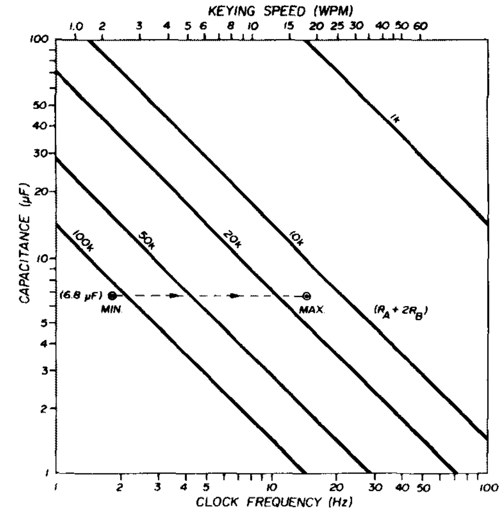


fig. 2. Capacitance, C, and resistance, (R_A + R_B), required to obtain desired keying speeds.

through a normally closed switch, which is bypassed by a 0.01 μF capacitor. When the clear switch is depressed, a 0.1 μF capacitor discharges into the reset input of the 7473 that controls the address cycle. This ensures that not only will the remainder of the address cycle during which the switch was operated be cleared, but that a new cycle will be started and cleared. In the write mode, complete memory erasure is provided.

transmitter keying and sidetone

A relay output to key the transmitter can be used to replace the 2N4888 keying transistor shown in the original circuit (see fig. **6**). The 5-volt reed relay, which is similar to that provided by Electronics Applications Company part no. 1A5AH,* provides excellent keying even at speeds above 35 wpm.

An improved keying monitor to replace the 7413 NAND Schmitt trigger[5] is also shown in fig. **6**. This

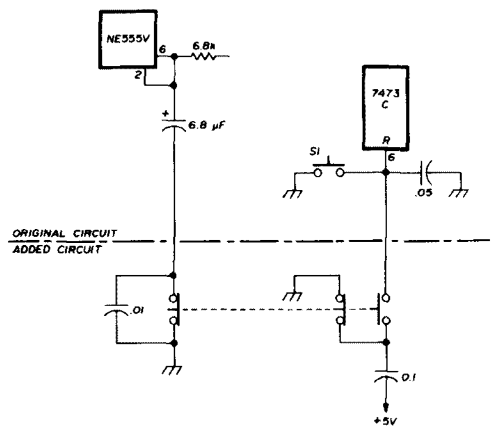


fig. 3. Addition of clear switch to clear keyer memory during read or write mode (contributed by K3DW).

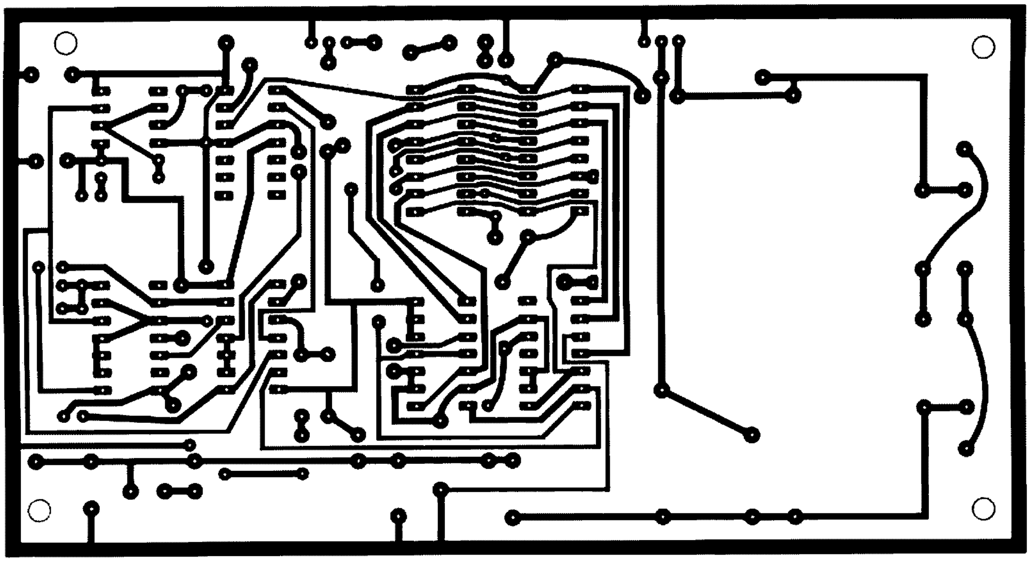
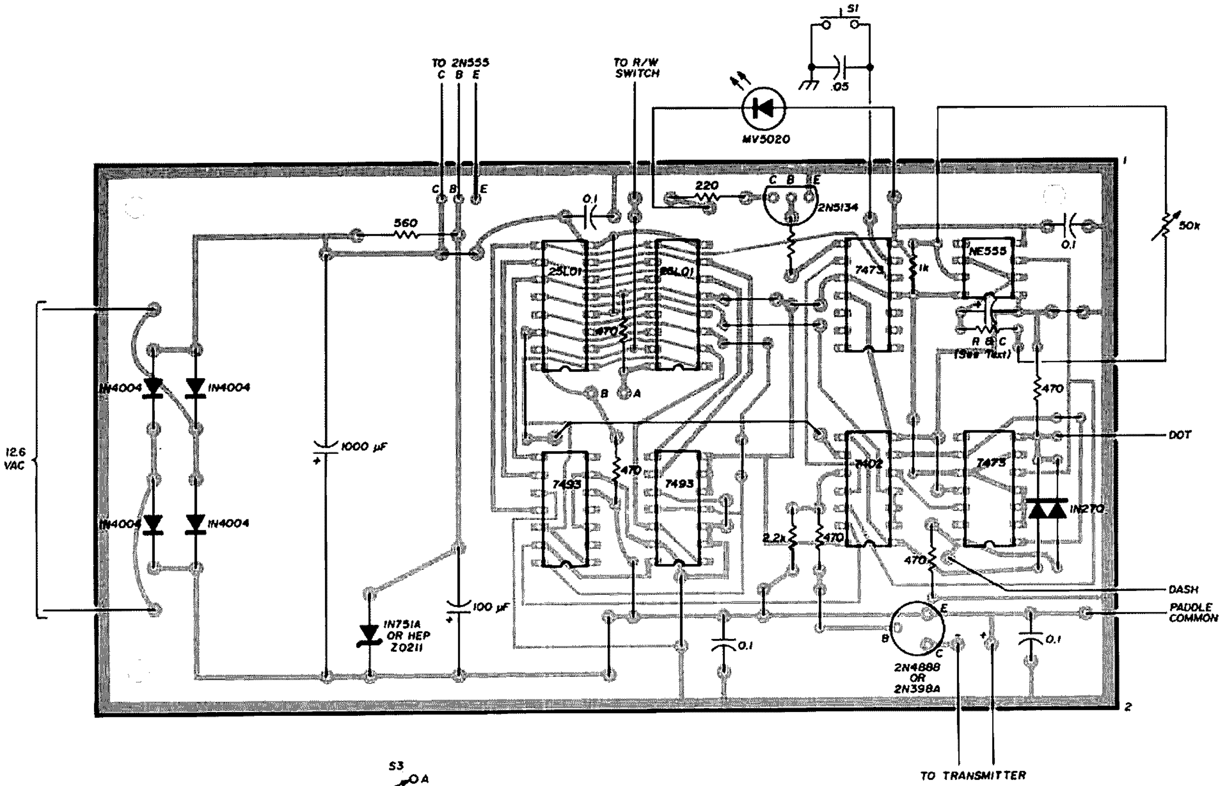


fig. 4. Full-size etched circuit board layout. Component placement is shown in fig. 5.



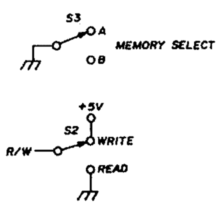


fig. 5. Top view, component side of PC board, with external connections.

circuit uses a NE555 wired as an astable multivibrator similar to a circuit used by WA5TRS.[6] If you don't want the added expense of the 500k pot, a resistor of about 150k ohms should provide a pitch pleasing to the ear, with the components shown.

construction notes

The printed-circuit board must be insulated from the metal cabinet by short standoff insulators. Also, if the keying paddle input jack is not insulated from the chassis, the PC board must be insulated from the cabinet. If you ground the board, insulate the input jack. In either case the keyer output jack should be insulated.

summary

All the troubleshooting concepts mentioned resulted from approximately 170 manhours debugging this keyer *after* it was assembled. Troubleshooting was done with a four-channel storage oscilloscope and a lot of patience.

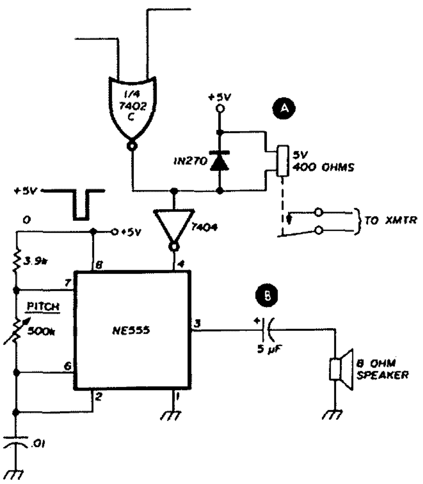


fig. 6. Alternative method for keying transmitter using a reed switch, (A). A simple keying monitor is shown in (B).

If you don't have access to this equipment, this article will be of use to you. You might want to include the additional memories described in reference 5.

*Electronics Applications Co., 2213 Edwards Avenue, South El Monte, California 91733.

references

1. Michael Gordon, WB9FHC, "Electronic Keyer with Random-Access Memory," *ham radio*, October, 1973, page 6.
2. Howard M. Berlin, K3NEZ, "Memory Keyer," (Comments), *ham radio*, December, 1974, page 58.
3. *The ARRL Radio Amateur's Handbook*, ARRL, Newington, Connecticut, 1973 Edition, page 356.
4. James M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.
5. Howard M. Berlin, K3NEZ, "Increased Flexibility for the Memory Keyer," *ham radio*, March, 1975, page 64.
6. Joe Buswell, WA5TRS, "Simple Integrated-Circuit Electronic Keyers," *ham radio*, March, 1973, page 38.

ham radio

CIRCUITS & TECHNIQUES

Ed Noll
W3FQJ

audio-power integrated circuits

Audio-power ICs are available in the 5-watt and higher ranges for many applications. Some include an integrally designed heatsink as part of the package (fig. 1). These devices are convenient for making inboard or outboard amplifiers when you need some additional audio punch. Most will drive 8- and 16-ohm speakers. For QRP work they can be used as a complete speech amplifier/demodulator for a-m, ssb, and fm. A modulation transformer can be added to match their low-impedance output to the transmitter. At the QRPP level, these ICs can be used as a single-module class-AB or class-B a-m modulator.

The RCA CA3131 and CA3132 (fig. 2) are two audio-power ICs that include preamps, power amplifier, and integral heatsink. The CA3131 has an internal feedback network that maintains an overall gain of approximately 48 dB. The CA3132 has no feedback network but has facilities for connecting one externally, depending on specific application. In this case the external feedback network usually connects between terminals 6 and 16. The package is a 16-pin dual-inline with the four center pins removed.

fig. 1. Sinclair IC-12 audio power IC provides up to 6 watts power output into an 8-ohm load. Voltage gain is about 250; input impedance is 250k.

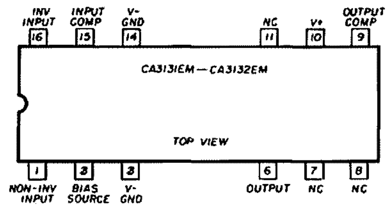


fig. 2. Pin-out diagram of the RCA CA3131 and CA3132 audio power ICs. The CA3131 has an internal feedback network that maintains overall gain at about 48 dB. The CA3132 has no internal feedback but one can be connected externally (see text).

Power output is 4 watts minimum and is typically 5 watts. Recommended supply voltage is 24 volts dc. The load can be either 8 or 16 ohms, with 8 ohms providing higher output. Zero-signal supply current is only 10 mA — certainly a favorable attribute for solar- and battery-power applications. Inverting and noninverting inputs are included. Output is single-ended; minimum input impedance is 200k but typically 1 megohm.

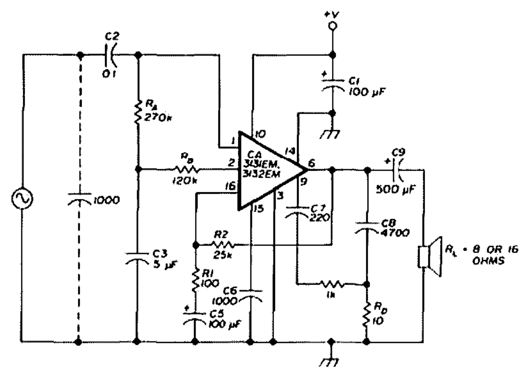


fig. 3. Five-watt audio power amplifier based on the RCA CA-3131/3132. The 1000 pF capacitor marked with an asterisk is required if the input has an open circuit.

A complete schematic including external components is shown in fig. 3. The audio signal is applied to the noninverting input, terminal 1, through C2. Input biasing is by R_A and R_B. R_B and C3 filter any ac ripple from the supply voltage line. As mentioned, the input impedance is high; therefore in a practical circuit the input impedance is largely set by the ohmic value of R_A.

Filter capacitor C1, an electrolytic, should be placed as near as possible to terminal 10. C6 sets a 46 dB closed-loop gain point at 200 kHz. C7 ensures equal gain characteristics on positive and negative signal swings. C9 sets the amplifier low-frequency response.

R1 and R2 are a part of the feedback network and need only be inserted when the CA3132 is used. C8 compensates for speaker inductance, with R_D limiting any current surge. Closed-loop gain equals the ratio (R1 + R2)/R1. The low-frequency 3-dB-down point occurs when C5 reactance equals the ohmic value of R1.

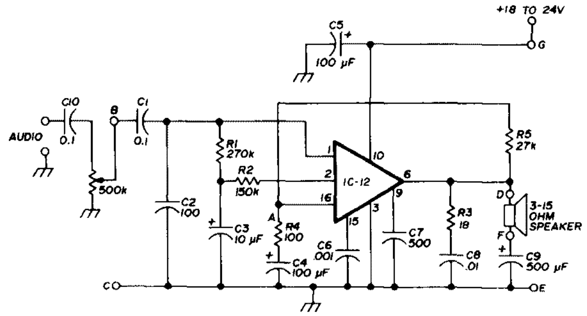


fig. 4. Five-watt audio power amplifier featuring the Sinclair IC-12 audio power IC. Zero-signal supply current is a low 10 mA.

The British Sinclair IC-12* (fig. 1) has a power output up to 6 watts with a 30 mV input when its output is connected to an 8-ohm load. Permissible output load impedance is 3 to 15 ohms. Voltage gain is approximately 250; input impedance, 250k. Zero-signal supply current is a low 10 mA. I've used the circuit of fig. 4 successfully for many receiver and modulator applications. Again, a modulation transformer can be substituted for the speaker if the modulated system reflects a 3 to 15 ohm impedance.

One of the more unusual applications of the IC-12 was in a broadcast-band receiver (fig. 5) using a phase-locked loop (PLL) detector. The PLL output was connected directly to the IC-12 audio-input terminals (fig. 4). Good performance and acceptable selectivity were obtained.

External parts values can be selected according to desired performance. For example, C1 (fig. 4) can be used to control bass rolloff. The 3-dB-down point is located approximately at the frequency at which C1

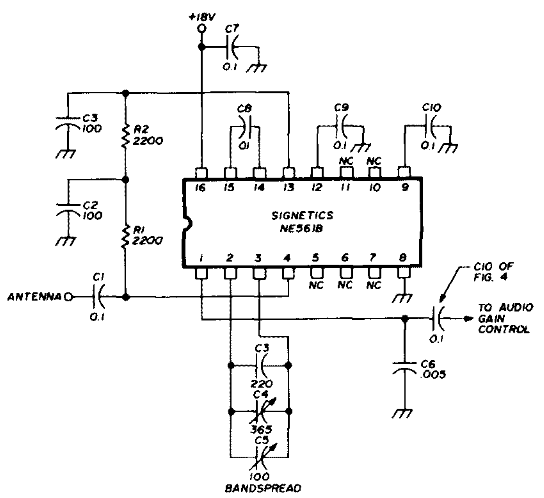


fig. 5. Phase-locked loop a-m detector which uses the Signetics NE561B. In a commercial broadcast-band receiver this detector was used with the IC audio power amplifier of fig. 4.

reactance equals the ohmic value of R1. C2 is needed only if the input signal source is from a very high impedance. C3 ensures good power-supply ripple rejection and low-frequency stability. Low-frequency rolloff is also influenced by R4 and C4. Rolloff frequency is that frequency at which reactance equals resistance. C5 is the power supply filter. High-frequency performance is influenced by C6 and C7, with C7 having its greater influence on the negative-swinging excursions. The value shown for C6 ensures low distortion up to 50 kHz. As in the previous schematic, fig. 3, C8 and R3 compensate for loudspeaker inductance, while C9 can be used to limit the bass response. R4 and R5 set the voltage gain:

$$V_G = \frac{R4 + R5}{R4}$$

The value of R5 can be increased to bring up the gain. For example, with a value of 470, the gain is 5000. In this case the input signal need only be 1 millivolt to produce rated output, but distortion is higher and careful layout is important to minimize stray feedback.

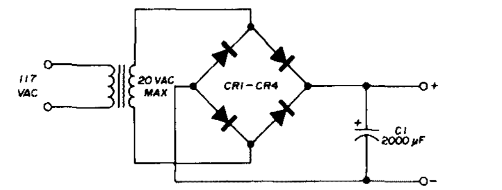


fig. 6. Suggested power supply. The diodes should be rated at 30 PIV, 1 amp. Transformer secondary voltage above 20 volts ac is not recommended.

However, if only a limited increase in gain is desired, the ohmic value of R5 can be increased gradually to meet specific needs and perhaps need not be increased to the point at which instability becomes a problem. If battery operation is not desired, a simple power supply can be built around a filament transformer as in fig. 6. Use a solid-state bridge rectifier, with each diode having a rating of at least 1 ampere and 30 volts PIV. A secondary voltage above 20 volts ac is not recommended.

multimode detector

The Plessey SL624C IC can be used to detect a-m, fm, ssb or CW signals. In ssb and CW reception it functions as a product detector with built-in oscillator. Operation as a quadrature detector recovers fm while a-m signals are demodulated with a synchronous detector. As an a-m detector, the SL624C is capable of rejecting broadband i-f noise as compared to a conventional envelope detector. The SL624C has been designed specifically for use in mobile, hf, and vhf transceivers. With a suitable circuit arrangement it can also be used to demodulate fm broadcast or TV audio signals.

The SL624C IC is shown in fig. 7. At left is an audio amplifier with input at pin 1 and outputs at pins 15 and

*Available from Audionics, 8600 NE Sandy Boulevard, Portland, Oregon 97220.

16. This amplifier has a gain of 12 dB. Included also is a limiting amplifier with input at pins 3 and 4 and output at pins 6 and 7. This amplifier operates up to 30 MHz and starts limiting with a 100-mV input level. Loop gain is about 70 dB. The limiting amplifier can be operated as

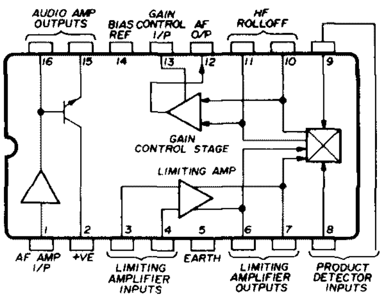


fig. 7. Plessey SL624C multimode detector can be used to detect a-m, fm, ssb or CW signals. Practical detector circuits using this IC are shown in fig. 8.

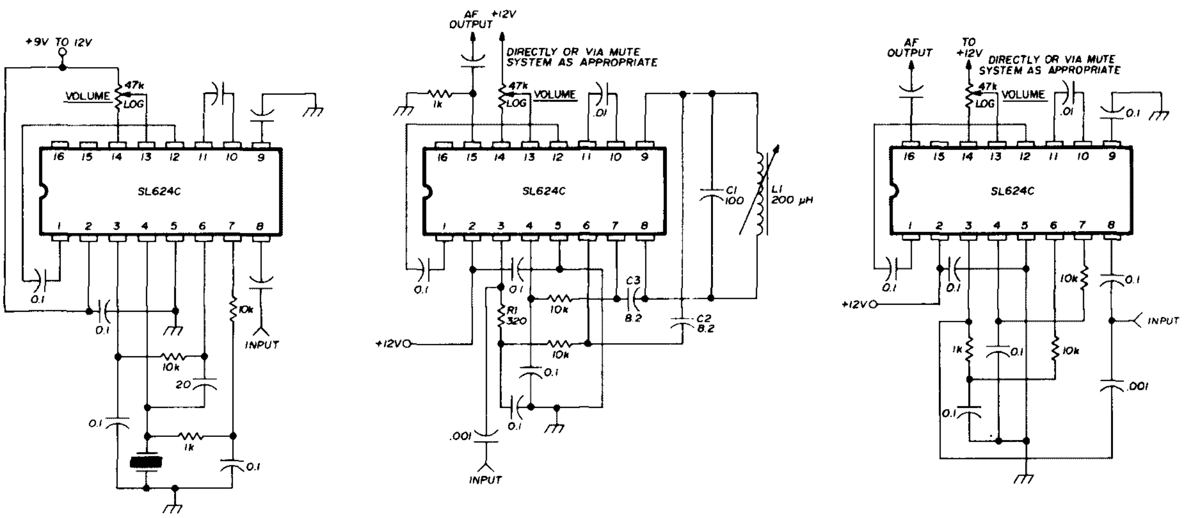
a beat-frequency oscillator as in the demodulation of an ssb signal.

The detector is a double-balanced modulator (like the Plessey SL640C). The limiting amplifier output is applied to the detector. The detector output connects to the audio gain control stage input. This gain can be regulated with a control voltage at pin 13. Audio output is removed from the gain stage and made available at pin 12. If desired, the output can be muted by connecting pin 13 to ground with a switch or electronically with a squelch circuit.

Practical detector circuits are given in **fig. 8**. In the sideband demodulation mode, the limiting amplifier is operated as a crystal oscillator. Its output is applied to the detector. Note the external crystal and the connection to pins 6 and 7. The sideband signal to be demodulated is applied through a coupling capacitor to pin 8. After passing through the gain control stage, the recovered audio is removed at pin 12 and applied through the 0.1 μF capacitor to the audio amplifier input through pin 1. Audio can be taken at either pin 15 or 16. The signal input requirement is 50 mV maximum, but good performance at a lower audio level can be obtained with an input as low as 5 mV.

In the synchronous detection of an a-m signal, input is applied to the detector through pin 8. Signal is also applied into the limiting amplifier through pin 3. In the limiting amplifier, the carrier is separated from the modulation and is used to generate a demodulating carrier component, which is applied to the detector and used to demodulate the incoming a-m signal. Adequate signal must be applied to permit limiting during modulation troughs to avoid distortion. The input signal should be 5 to 50 mV. An external agc system is recommended for this detection mode.

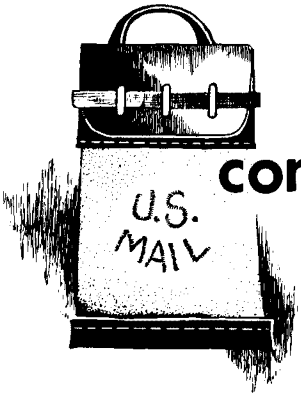
In fm detection, the signal is applied to the limiting amplifier input at pin 3, then through a phase-shift network to the detector input. Also the quadrature component is applied to the detector input through pins 8 and 9. Note the resonant circuit C1-L1. The detector output is proportional to the relative phase of the two inputs, with the quadrature component (which does not deviate) serving as a reference phase. The recommended input signal should be at least 200 μV, although demodulation occurs with an input signal as low as 100 μV. The only adjustment required is that of the phase-shift circuit. An external squelch circuit is used to reduce high-level noise when no signal is being received.



A SELF-OSCILLATING SSB DETECTOR
B FM QUADRATURE DETECTOR
C SYNCHRONOUS A-M DETECTOR

fig. 8. Practical detector circuits using the Plessey 624C include a self-oscillating ssb detector (A), fm quadrature detector (B), and synchronous a-m detector (C).

ham radio



comments

microstripline preamplifiers

Dear HR:

WA6UAM's article on "Microstrip Preamplifiers for 1296 MHz,"[1] with a few exceptions explained below, is an excellent article. Having worked with stripline for several years, especially in development of the TIROS-ESSA antenna matching circuitry, I can attest to the value of such a practical construction article for the uhfer. It was also very timely, as more and more amateurs are starting to use stripline techniques to build uhf equipment.

However, in the design section of the article, several unfortunate errors and contradictions appear in the treatment of the S-parameter reflection coefficients and impedances, which are confusing and misleading, even to one who is familiar with S-parameter techniques. The confusion begins in the first paragraph on page 22, where the author states that complex impedances are generally shown in polar form, but can be converted to rectangular form through use of the Smith chart, as per instructions in the caption of fig. 12. The inference is quite clear that the conversion intended is between the polar and rectangular forms of an equivalent value of *impedance*. However, it is not *impedance* which is being converted, and furthermore, the Smith chart cannot perform this type of conversion. Therefore, the inference is incorrect.

The confusion is compounded in the next paragraph, where it is stated that table 1 lists complex *impedances* in both polar and rectangular forms, while in the table itself both the polar and rectangular forms are stated to be *reflection coefficients*. This contradiction needs clarification, and the statements emphasize the previous, erroneous inference that the associated values appearing in polar and rectangular form in the table are numerically equivalent, while in fact they are not.

The confusion can be easily cleared up as follows: First, it is evident that the author is randomly interchanging reflection coefficient and impedance, confusing the polar-form reflection coefficient with the polar-form equivalent of the rectangular-form impedance. The two are not the same!

Impedance, $Z = E/I$, describes the relation between voltage and current in a circuit. Reflection coefficient, ρ, on the other hand, is the relationship between two voltages (the reflected and the incident) in a circuit containing *two* impedances at a junction, or two currents in the same circuit:

$$\overline{\rho} = \frac{E_{reflected}}{E_{incident}} = -\frac{I_{reflected}}{I_{incident}}$$

Accordingly, to clarify the first paragraph on page 22 of WA6UAM's article, the phrase "complex *impedances* in polar form . . . " is a mistatement which should be changed to read "complex *reflection coefficients* are generally shown in polar form, which can be converted to *impedance* in rectangular notation (R ± jX) on a Smith chart as indicated in fig. 12" (after the caption of fig. 12 is also corrected).

Second, the complex numbers appearing in *polar* form in table 1 are reflection coefficients, and the rows containing the polar-form values should be so labelled. Third, the complex numbers appearing in *rectangular* form in table 1 are the *impedances* which will give rise to the accompanying value of reflection when terminating a line or source having an impedance of 50 ohms. In other words, taking an example from the second HP-25826E column, the 12.5 + j0.5 value is *not* the rectangular equivalent of the polar value 0.61 $\angle 178^\circ$, but is the complex impedance which will *yield* the complex reflection coefficient $\overline{\rho} = 0.61 \angle 178^\circ$ when the impedance 12.5 + j0.5 terminates a 50-ohm line or source. The rows containing complex numbers in the rectangular form should therefore be specifically labelled *impedance* S_{11} or S_{22}, as appropriate. Proof that the rectangular-form impedance is not equivalent to the listed polar value is further shown by the fact that the polar equivalent of the impedance 12.5 + j0.5 is actually 12.51 $\angle 2.29^\circ$, and *not* 0.61 $\angle 178^\circ$.

Fourth, as constructed in figs. 9, 10, 11 and 13, the graphs containing the S_{11} and S_{22} plots should be labelled *impedance*, not "reflection coefficient" because the only loci-identifying coordinates in the graphs are the resistance and reactance circles. The S-parameter graphs in the Hewlett-Packard design catalog[2] from which the figures in the article were taken contain *two* sets of coordinates by which the loci may be identified: resistance- and reactance-circle coordinates to identify the loci as impedances, *and* radial magnitude and angle coordinates to identify the loci as reflection coefficients. Thus the user could use whichever set of coordinates he desired to read the loci as impedances or reflection coefficients.

It is apparent in unravelling all this confusion that a misunderstanding also exists concerning the basic functions of the Smith chart. The function which the Smith chart is really performing in fig. 12 is the conversion from the complex

1. H. Paul Shuch, WA6UAM, "Microstropline Preamplifiers for 1296 MHz," *ham radio*, April, 1975, page 12.

2. "Diode and Transistor Designers Catalog," Hewlett-Packard, May, 1974.

reflection coefficient in the polar form to the *normalized impedance* in the rectangular form. The magnitude (radius) and angle $0.8\ \angle{-50^\circ}$ in fig. 12 define a specific point in *reflection-coefficient* coordinates of the chart, while normalized impedance is found at this same point where the r and x impedance coordinates of 0.6 and 2.0 intersect, respectively. It cannot be emphasized too strongly that the chart is *not* converting impedance in the polar form to its equivalent impedance in the rectangular form.

Polar-to-rectangular conversion of equivalent impedances is relatively simple to calculate using the Pythagorean theorem. However, conversions between reflection coefficient and impedance are more difficult to calculate, hence the Smith chart is used to simplify reflection-to-impedance conversions. As a point of interest, polar-to-rectangular impedance conversions *can* be performed with an overlay combination of Smith and Carter charts having the same diameters (the Carter chart has impedance coordinates arranged to identify impedance in polar form). With the Smith-Carter overlay the user may enter the Smith chart in rectangular form and the corresponding point on the Carter chart is the polar-form equivalent. As a further point of interest, here is the expression for calculating the conversion from a complex reflection coefficient $\overline{\rho}$ to *normalized* impedance:

$$\frac{Z}{Z_c} = \frac{R+jX}{Z_c} = \frac{1+\overline{\rho}}{1-\overline{\rho}} = \frac{1+\rho\angle\theta}{1-\rho\angle\theta}$$

$$= \frac{1+\rho\cos\theta + j\rho\sin\theta}{1-\rho\cos\theta - j\rho\sin\theta}$$

Going in the opposite direction, to determine the reflection set up by a given complex impedance loading a line of impedance Z_c, we have

$$\overline{\rho} = \rho\angle\theta = \frac{R+jX-Z_c}{R+jX+Z_c}$$

Two additional errors of lesser importance are, first, on page 25 at the beginning of column 2, the shunt equivalent value of the series impedance 40 + j25 ohms should be changed from 34.8 +j55.6 ohms, to read *55.6 + j89* ohms. And second, the NEC VO21 column of table 1, the reactance −j38.5 in the parallel-circuit input impedance should be changed to indicate a *positive* reactance.

As a final point of interest, in 1953 the American Standards Association (ASA) adopted the Greek letter rho, ρ as the symbol to represent reflection coefficient, and many textbook and periodical publishers, as well as manufacturers of S-parameter measuring instrumentation, conformed. Prior to 1953, ρ was often used to indicate swr, while gamma, Γ and k were used interchangeably to represent reflection. It would be interesting to know why the people at Hewlett-Packard who produce solid-state components continue to use Γ, while those who produce the instruction manuals for their impedance and S-parameter measuring equipment are using ρ.

Walt Maxwell, W2DU
Dayton, New Jersey

W2DU has raised a valid point with regard to the rather loose terminology which I used in my recent article, and I concede that reflection coefficient and impedance are not synonomous, although they are related.

Several readers have questioned my failure to consider the transistor's transfer coefficient in calculating the matching networks. Actually, my simplistic design method, which ignores S_{12} in particular, results in a minute matching error which may be compensated by adjusting the trimmer capacitors at the input and output of the preamplifier.

For the benefit of those readers who have inquired about Rollett's stability factor, I should mention that K calculates to greater than unity for all transistor/bias combinations presented in the original article so the amplifiers are unconditionally stable. Nevertheless, I caution the builder to treat them as though they were not. That is, do not apply power until the amplifier is properly terminated in an antenna (or dummy load) and a converter

H. Paul Shuch, WA6UAM

HP-65 oscar tracking program

Dear HR:

A program, written for the Hewlett-Packard HP-65 programmable calculator is available for computing both azimuthal and elevation coordinates for tracking either OSCAR 6 or 7. Once the satellite orbital data and individual station positional coordinates are inputed and stored, the routine will compute az/el antenna pointing coordinates for any number of arbitrary, specified times following the ascending node. Az/el coordinates computed with this routine for both ascending and descending passes agree favorably with the results of a FORTRAN program run on a CDC-3800 computer and with actual observed satellite trajectories. The program can be stored on a single HP-65 magnetic card. Documented copies of the program will be forwarded upon request and receipt of a self-addressed, stamped envelope; if a blank magnetic card is included with the request, a copy of the program itself will be sent.

Earl F. Skelton, WA3THD
Washington, DC

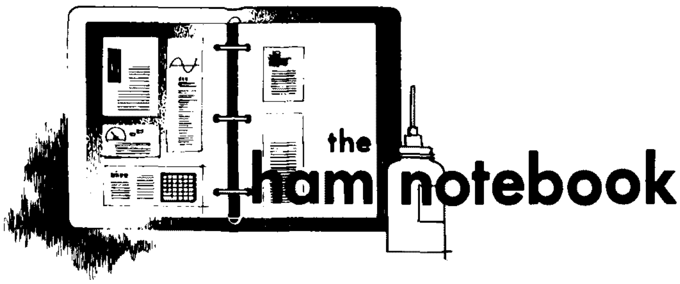
lower telephone rates

Dear HR:

I am sure many of the readers of *ham radio* have seen the recent ads run by the telephone company depicting the new, low long-distance rates. For the minimum of one minute for 56¢ (at times, even less, depending on the distance) one may call coast to coast. In the evenings, from Sunday to Friday, 8 pm to 11 pm, a one-minute telephone call costs 36¢ or less. For nights and weekends, every night from 11 pm to 8 am and Saturdays and Sundays, the first minute is only 22¢.

It occurs to me that many hams who wish to communicate with another ham anywhere in the U.S., may alert the other party by placing a one-minute call, which would be all the time necessary to convey information as to frequency and a scheduled time. Previously, minimum rates were for 3 minutes and at triple the price. It's a good point to keep in mind when wishing to get another station on the air.

David Greene, W2IAO
West Orange, New Jersey



versatile audio oscillator

Here's a versatile audio oscillator which can be put to many uses including an audible logic indicator, sidetone oscillator, code-practice oscillator, square-wave signal generator and many others.

In the circuit of fig. 1 transistors Q2 and Q3 are arranged as a basic collector-coupled astable multivibrator; power is taken from the collector of Q1 which acts as a switch for Q2 and Q3. With S1 closed and S2 open, Q1 is cut off and the B1 battery potential is furnished to Q2-Q3 through R1. With both S1 and S2 closed, Q1 is saturated and its collector potential drops to near ground; therefore, no voltage is available for Q2-Q3 and oscillation ceases.

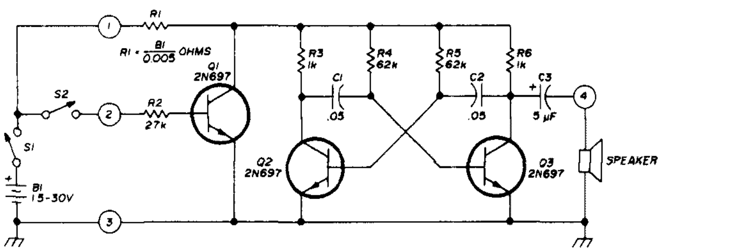


fig. 1. Versatile audio oscillator circuit which may be used as an automobile headlight indicator, audible logic indicator, sidetone oscillator or square-wave signal generator. Oscillation frequency can be varied by changing R4 or R5.

The frequency of oscillation is essentially independent of the B1 supply voltage and is determined by:

$$f = \frac{1}{0.69(C1 \cdot R5 + C2 \cdot R4)}$$

If R4 = R5 and C1 = C2, then the output will be a symmetrical square wave. The frequency of oscillation can be varied by changing the value of either R4 or R5 or both; however, if R4 and R5 are not changed a like amount, output symmetry will be lost. With the circuit values shown, a 100k pot in series with a 20k resistor could be substituted for R5.

The oscillator output is taken from the Q3 collector via C3. The size of C3 has a marked effect on output volume when a low-impedance load, such as a speaker, is used. Values of 2 µF or larger are quite satisfactory for all impedance loads and will furnish ample audio volume. If only high-impedance loads are used such as 2k headphones, a 0.05 µF disc capacitor will provide adequate audio coupling. If a better impedance match and slightly more volume are desired, an audio output transformer may be used (fig. 2).

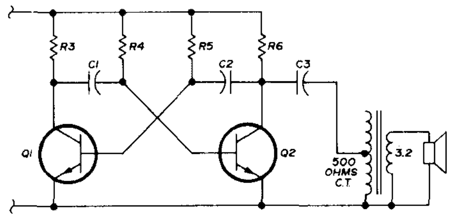
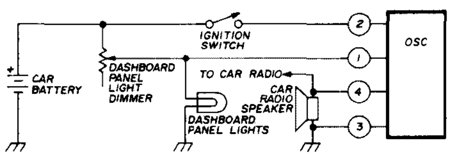


fig. 2. Slightly more audio output can be obtained from the oscillator of fig. 1 by using an audio output transformer.

When used as an automobile headlight reminder (with a negative-ground car) connect the circuit as follows:



Power for the oscillator is derived from the dashboard panel lights, which are turned on simultaneously with either the parking lights or headlights. If the ignition key is turned on, Q1 saturates and disables Q2-Q3; with the ignition off Q1 is cut off and the Q1 collector voltage rises, providing power to Q2-Q3. The audio output may be connected directly to the car radio speaker voice coil high side without affecting car radio operation.

By connecting the oscillator port 1 to the panel lights the oscillator may, if desired, be purposely disabled with

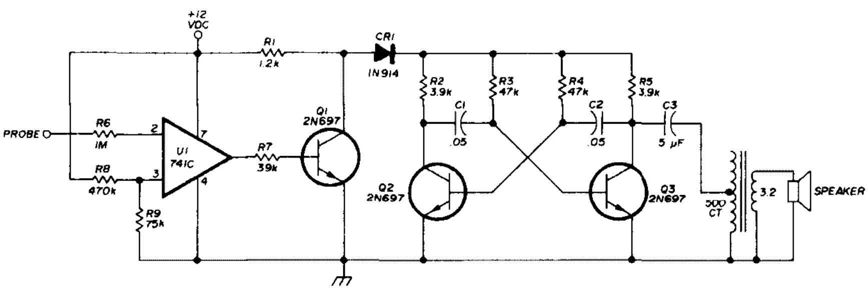
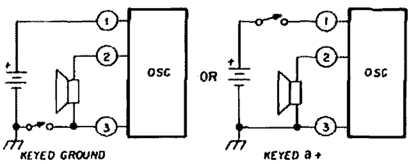


fig. 3. Using the audio oscillator as an audible logic indicator. Oscillator is isolated from the logic by op amp U1 which is wired as a Schmitt trigger.

the ignition off and lights on merely by dimming the dashboard panel lights. Current drain of the oscillator on the car battery is virtually negligible. The oscillator may be permanently wired to and powered from the existing dashboard controls without requiring additional controls or switches.

The entire printed-circuit board can be wrapped with electrical insulating tape and strapped to any convenient location under the dashboard out of sight, or mounted in a small Minibox. For connection to the car's electrical system, the proper leads can be easily located with a voltmeter or VOM; once located, simply splice in the appropriate oscillator lead, solder, and wrap the joint with electrical tape.

When used as a sidetone oscillator or code practice oscillator, connect as follows:



In the above configurations, Q1 and R2 may be eliminated, if desired. The entire oscillator may be constructed on a PC board measuring only 1-1/8 by 3/4 inch (29 by 19mm) if TO-92 transistors are used. For TO-5 transistors the board is slightly larger, 1¼ by 7/8 inch (32 by 22mm). Height of the board with components is about ½ inch (13mm). Since the circuit is very simple, point-to-point wiring on terminal strips is another alternative if automobile installation is not intended.

Layout of components is not critical, nor is selection of Q1-Q3. Although 2N697s are specified, unmarked npn transistors from surplus

computer boards were used for the fifteen or so automobile headlight units I've built so far; all worked as intended the first time.

For the IC enthusiast, Q1-Q3 can be individual transistors in an array such as the CA3018 (TO-5 case) or the CA3046 (14-pin DIP). However, if ICs are used, a 1N914 diode will be necessary from the Q1 collector to R3; otherwise, it may be omitted.

When the oscillator is used as an audible logic indicator, or audible logic-state indicator, additional isolation of the oscillator from the probed circuit element should be provided to prevent loading the logic circuit. A high-impedance input op-amp is ideal for this application. Fig. 3 shows the circuit.

The op-amp is configured as a poor-man's Schmitt trigger; i.e., a fairly rapid output transition occurs at a specific preset input voltage level by omitting the usual feedback resistor between pins 6 and 2. The op-amp acts simply as a very-high-input-impedance inverter with virtually no hysteresis about the preset transition reference voltage level appearing at the non-inverting input, pin 3. This reference voltage is easily provided by the resistive divider network R8-R9.

Since a TTL-compatible logic probe was desired, the reference level was set for +1.6 volts. The zero logic state *maximum* voltage for the SN7400 series TTL ICs is about 0.8 volt; the minimum 1 logic level is about +2.4 volts. The +1.6 volt reference level is an arbitrary selection between the two TTL logic levels. When the probe input voltage is below +1.6 volt the op-amp output is approximately 10.5 volts, which saturates Q1 and disables Q2-Q3; when the probe voltage is above +1.6 volt, the U1 output is about 2 volts, which cuts off Q1, and power is supplied to Q2-Q3. R7 must be selected to allow cutoff of Q1 when the U1 output is low, and permit saturation of Q1 when the U1 output is high.

These are just a few of the possible applications for this handy and inexpensive oscillator; further applications are left to the ingenuity and imagination of the reader.

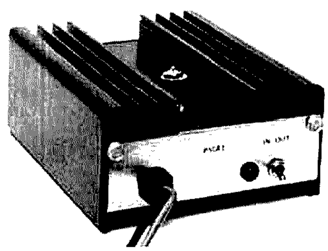
Howard F. Batie, W7BBX

NEW products

two-meter fm transceiver

The new products announcement of Standard's new Horizon 2, a 12-channel, 25-watt vhf fm transceiver, in the November issue of *ham radio* contained a typographical error: the correct amateur price is $295.00 Contact your local dealer for further details.

vhf fm power amplifiers



The new M-Tech P50A1 vhf power amplifier is designed specifically for amateurs with low power two-meter fm transceivers or hand-held units – 1 to 3 watts input will deliver 40 to 65 watts output. The P50A1 is designed for operation with a 13.6-volt power supply (8 amps) and is rated for an 85% duty cycle. The unit includes COR switching with an LED indicator and a spurious output filter, and is priced at $139 postpaid.

Other 144-MHz fm power amplifiers in the M-Tech line include the P15A1 (1-3 watts input, 12-25 watts output, 100% duty cycle) which features solid-state switching and is priced at $59; the P50A10C (2-18 watts input, 14-60 watts output, 100% duty cycle), $98; the P100A10 (5-12 watts input, 60-100 watts output, 85% duty cycle), $198; the P100A20 (18-35 watts input, 80-100 watts output, 85% duty cycle), $155; and the P100A5 (2-5 watts input, 40-100 watts output, 85% duty cycle), $198. All amplifiers are vswr protected for *any* load, include a reverse current protection circuit, use microstrip inductors for stability, and carry a 1-year factory warranty. All amplifiers except the P15A1 feature COR switching with an LED indicator and a spurious output filter.

M-Tech also manufactures two solid-state power amplifiers for the 220-MHz band, the P30A1-220 (1-3 watts input, 30-45 watts output, 85% duty cycle) and the P30A10-220 (2-18 watts input, 12-40 watts output, 100% duty cycle). Both of these units feature COR switching with LED indicator.

For more information on M-Tech's *Quality Emphasis Line* of vhf-fm power amplifiers, write to M-Tech Engineering, Inc., Box C, Springfield, Virginia 22151, or use *check-off* on page 102.

random-wire antenna tuner

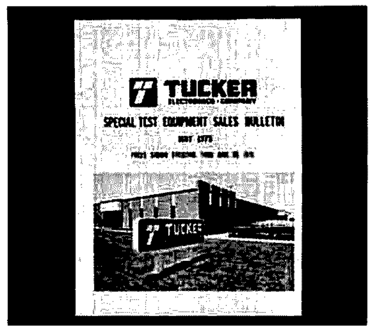
If you like portable operation and want to get on the air with the least amount of trouble, a random-length wire antenna is hard to beat. You'll need a tuner for the random wire, and SST Electronics has the answer with the SST T-1. The SST T-1 tunes from 80 through 10 meters and handles 200 watts. It matches the low-impedance output of your transmitter and the low-impedance input of your receiver to the high impedance of a random-length wire antenna. Simple and foolproof design features an LC circuit and neon-bulb tune-up indicator. It's compact, only 3 by 4¼ by 2-3/8 inches (7.6 by 11x1cm). The SST T-1 sells for $24.95 postpaid and is guaranteed for 90 days against defects in parts and workmanship. For more information, write SST Electronics, P. O. Box 1, Lawndale, California 90260, or use *check-off* on page 102.

power-line monitor

A new compact high-low power-line monitor with a convenient swivel plug for use directly in an ac outlet or through a standard multi-socket cube is now available from RCA. This small inexpensive test instrument is an ideal tool for every amateur's toolbox and reads from 50 to 150 volts ac (true rms), 50-60 Hz with a plus or minus 5 per cent accuracy. Circular in shape, the new monitor is only two inches in diameter and one inch deep (5x2.5cm), and weighs only three ounces (85g).

The RCA WV-548A Hi/Low power line monitor is priced at $9.95. For additional information on RCA Electronic Instruments contact RCA Distributor and Special Products Division, 2000 Clements Bridge Road, Deptford, New Jersey 08096, or use *check-off* on page 102.

test equipment



The 24-page Tucker Electronics Sales Bulletin lists a wide variety of reconditioned test equipment as well as a dozen different lines of new instruments. Although the bulletin shown above was released in May, new sales bulletins are issued periodically. For your copy, write to Tucker Electronics Company, Post Office Box 1050, Garland, Texas 75040, or use *check-off* on page 102.

volt-ohm-milliammeter

The Triplett Corporation has introduced an unconventional type of volt-ohm-milliammeter that gives the user an "extra-chance" after misuse . . . and not a repair bill. This virtually indestructible test instrument has built-in protection against accidental high energy overload, is shock resistant to accidental drops up to a five foot (1.5m) height, is of modular construction so that it can be easily and quickly serviced in the field and has been designed to the most rigid safety standards to prevent any hazard of electrical shock to the user. Triplett has aptly named it the "Extra-Chance" model 60.

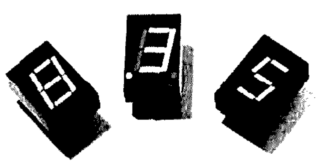
The new vom has no exposed metal parts, providing complete insulation of the instrument itself, special test leads for increased safety and a three-fuse testing system which greatly reduces fire and explosion hazard under misuse conditions. Two 48-inch (1.2m) long safety

test leads are supplied and connect to the control panel by special safety connectors.

A rugged case molded of black, high impact thermoplastic material in combination with a ruggedized suspension meter result in a vom that is virtually indestructible from accidental drops up to a five-foot (1.5m) height. The meter movement is protected by a diode module; fuses are used for normal overload conditions. A fuse plus two zener diodes are used to protect against high energy fault currents and protect the circuit up to 1000 volts. A separate, sealed battery compartment permits easy external access to batteries and fuses without having to remove other parts of the instrument.

A single range selector switch is used for selecting all 33 ac/dc voltage, ac output, resistance, dc current and decibel ranges from -20 to +52 dB plus the *off* and *test* positions. Accuracy on all dc and resistance ranges is ±2 per cent of full scale; ac accuracy is ±3 per cent of *full scale*. The Triplett model 60 (catalog number 3145) comes complete with a one-year parts and labor warrantee, safety test leads, batteries, spare fuses and instruction manual, and sells for $90. For additional information, write to the Triplett Corporation, Department, PR, Bluffton, Ohio 45817, or use *check-off* on page 102.

seven-segment displays



New high-efficiency solid-state numeric displays, as much as five times brighter than other displays at the same operating current, are now available from Hewlett-Packard. At one-fifth the current, they are equal in brightness to older displays. Their high brightness plus their 0.43 inch (11mm) height makes them ideal for applications in high ambient light conditions. Operating at currents as low as 3 mA, these large displays become practical for use in battery-powered portable instruments. They

may also be operated in the strobe mode at currents up to 60 milliamperes peak.

The HP models 5082-7650, -7660, and -7670 are red, yellow and green, common-anode, seven-segment displays with lefthand decimal point. LED chips are optically magnified to form evenly-lighted segments. For improved on-to-off contrast, the bodies of the displays are colored to match the appearance of the unlighted segments.

For more information, contact your local Hewlett-Packard Sales Office, or use *check-off* on page 102.

fet multimeter

A new pocket fet multimeter offering full vtvm ranges and a 10-megohm input, completely protected against overload, is now available from Hickok. Packaged in a rugged, pocket-size case with attached cover the Model 350 provides features which include 1 millivolt resolution on three easy-to-read mirrored scales plus dB and battery condition, high/low ohms ranges, and true autopolarity with a polarity indicator.

High impedance fet circuitry permits vtvm type ranges in this compact unit. Nine voltage ranges of 0.1 to 1000 volts and seven high/low ohms ranges from 100 ohms to 100 megohms center scale make the Model 350 a versatile service tool. One-year service can be expected from the two 9-volt transistor radio batteries. The Model 350 comes complete with two test leads and instruction manual.

For more information, contact Tom Hayden, Instrumentation & Controls Division, Hickok Electrical Instrument Company, 10514 Dupont Avenue, Cleveland, Ohio 44108, or use *check-off* on page 102.

More Details? CHECK—OFF Page 102

one dollar

ham radio magazine

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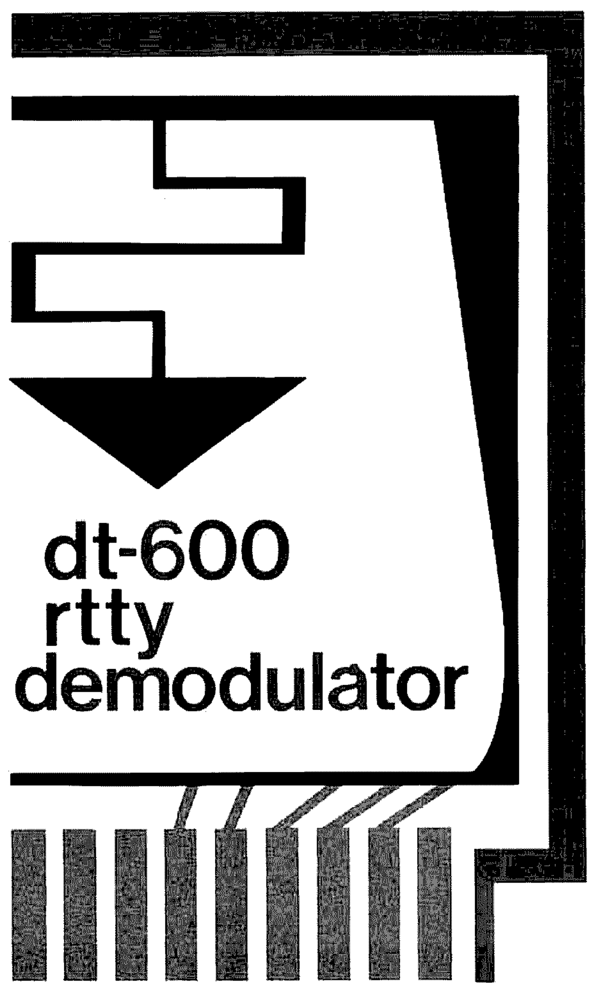

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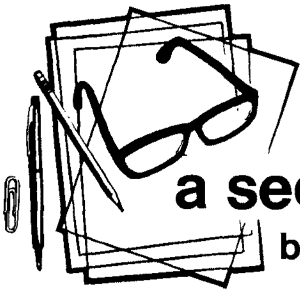


a second look

by Jim Fisk

With AMSAT-OSCAR 7 now well into its second year of continuous operation, it's becoming more and more apparent that an increasing number of amateur stations using mode B (the 432 to 144 MHz repeater) are using much more power than the recommended 100 watts *effective radiated power* (erp). When more than the recommended power is used, it swamps the agc circuit in the transponder, resulting in an excessive amount of current being drawn from the on-board power source.

In maximum sunlight the solar panel on the spacecraft can supply approximately 1 ampere of current; if more than 1 ampere is required to power the repeater it must be supplied by the on-board battery. However, the red-line limit on battery discharge current is about 1.2 ampere, and when amateurs using mode B run excessive power, the current drawn from the battery often *exceeds 2 amperes* when OSCAR 7 is in sunlight, and more than *3 amperes* when the satellite is in darkness.

At times this heavy current drain on the battery has caused the battery voltage to drop to the point where the under-voltage protection circuits have taken over. These circuits were designed to place the spacecraft in mode D (the discharge mode, both transponders turned off) when the battery voltage drops to 12.1 volts. The spacecraft systems have already switched to mode D a number of times, and some unexpected switches to mode A (the 144 to 29 MHz repeater) have also occurred.

Amateurs who are running more than the recommended 100 watts erp are conspicuous because their signals are much louder than the rest of the stations on the channel. If you tune across the mode B passband and note that certain stations in your local area are consistently much louder than others in the passband, please contact them directly, explain the adverse effects of excessive erp, and ask them to reduce power.

If they indicate that they're only running 100 watts *output*, ask them what they're using for an antenna – one-hundred watts of rf into a *single* KLM, Tilton, K2RIW or WØEYE Yagi produces an effective radiated power of 2000 to 3000 watts! Stations running 100 watts rf output into multiple Yagi arrays may have effective radiated powers of 8 kW or more. Some of the worse offenders appear to be a few 432-MHz EME operators who are not using 100 watts, but their kilowatt finals, and not with a single no-gain antenna but with their multiple Yagi arrays which have gains of 20 dBd or more, for an effective radiated power greater than 60 kilowatts. Their signals are brutally loud, but you can tell when they're on the air by simply monitoring telemetry channel 2B.

Since there may be some amateurs who don't understand the meaning of effective radiated power, following is a list of popular 432-MHz Yagis, and the rf power input for 100-watts effective radiated power for single, double and quadruple Yagi arrays:

antenna type	approx gain	single antenna	double array	quadruple array
KLM	15 dBd	3.2 W	1.6 W	0.9 W
Tilton	13 dBd	5.0 W	2.5 W	1.4 W
K2RIW	15 dBd	3.2 W	1.6 W	0.9 W
WØEYE	14 dBd	4.0 W	2.0 W	1.1 W

The approximate gain is that for a single antenna. A 3 dB increase in gain is assumed for a double array, and a 5.5 dB increase for a quadruple array. Note that 10 watts input to any of these antennas results in more than the recommended 100 watts erp – 10 watts into four properly phased Yagis produces an erp of 1000 watts or more.

These facts must be brought to the attention of mode B users who are abusing the recommended maximum 100-watt limit. Any assistance you can provide in reducing what has become a serious problem will be appreciated by other mode B users and AMSAT alike – continued abuse of the 100-watt power limit will most certainly shorten the useful life of the satellite.

Jim Fisk, W1DTY
editor-in-chief

presstop

AN ARRL OFFER to carry the brunt of the Amateur Radio WARC preparation effort was enthusiastically received at the December 12 meeting of the entire Working Group in Washington. League support outlined by ARRL General Manager, Dick Baldwin, will even extend to participation in various contributory activities such as CCIR meetings.

Overall WARC Timetable was reviewed by Chairman John Johnston and despite the apparent remoteness of 1979, it was obvious that the task of coordinating all services' frequency needs for the next two decades will require every bit of that time.

Amateur Frequencies Proposed by the various task forces include 160-200 kHz, 1715-2000 kHz, 3.5-4 MHz, 7-7.5 MHz, 10.1-10.6 MHz, 14-14.5 MHz, 18.1-18.6 MHz, 21-21.5 MHz, 24-24.5 MHz and 28-29.7 MHz in the HF spectrum, all on an exclusive basis. VHF frequency requests include 50-54, 144-148, 220-225, 420-450, and at least a portion of 890-942 MHz. Basic microwave allocations proposed don't differ much from present U.S. Amateur assignments, though a number of sub bands for satellite and experimental work were incorporated. How likely we are to get any of the proposed new bands or what problems we'll have in keeping or expanding those we have was debated at some length — though some services presently using a lot of the HF spectrum are planning to move to satellites, other services are eager to move into their slots, while pressures on VHF Amateur bands are well known and increasing.

HOAX DISTRESS SIGNAL showed up on 3804 kHz Christmas afternoon and tied up many Amateurs and Coast Guard people through the following afternoon. "WN8HOM" (a call unassigned by the FCC) claimed to be stranded on a 25 foot boat with 10 people on board that was disabled by engine failure near Pelee Island in western Lake Erie. WB9BWU called the Coast Guard in Detroit and the 170-foot Cutter "Mariposa" was dispatched to search for the lost vessel while a growing number of listeners throughout the U.S. monitored the frequency. WB9BWU and W8LIO became the relay stations between the "vessel in distress" and the "Mariposa" and Coast Guard land stations, all operating on 3804.

Possibility Of A Hoax was suspected fairly early in the adventure, but Coast Guard's philosophy is that all distress calls are genuine until proven otherwise. FCC monitors reported that WN8HOM's sporadic signals looked to them to originate from near Zanesville, Ohio, and listeners as far away as Florida reported them to be over S9 — certainly suspect for a Novice rig working from a small boat with failing batteries. Still the drama continued, complete with helicopter after improving weather permitted, until a Zanesville area Amateur reported the signals steady at S9+ from his location and then put his signals on 75 to let the FCC DFers confirm their earlier determination that the Zanesville area was the source. A few minutes later — at 2215 Z Friday afternoon — the Coast Guard called off the search.

AN AMATEUR'S ORIGINAL LICENSE — not a photocopy — will have to be submitted with his application for renewal under the terms of Docket 20672 released by the FCC in mid December. All Comments on the Notice of Proposed Rule Making must be filed by January 22, and Reply Comments are due February 2. This proposal apparently resulted from the discovery of a number of recent applications that included photocopies of licenses showing class not in agreement with FCC file information.

License Renewal Requirements for operating time and code proficiency were dropped in a "Christmas present" Report and Order adopted by the FCC. The relaxation, which became effective December 24, is a logical one since the requirements were essentially unenforceable.

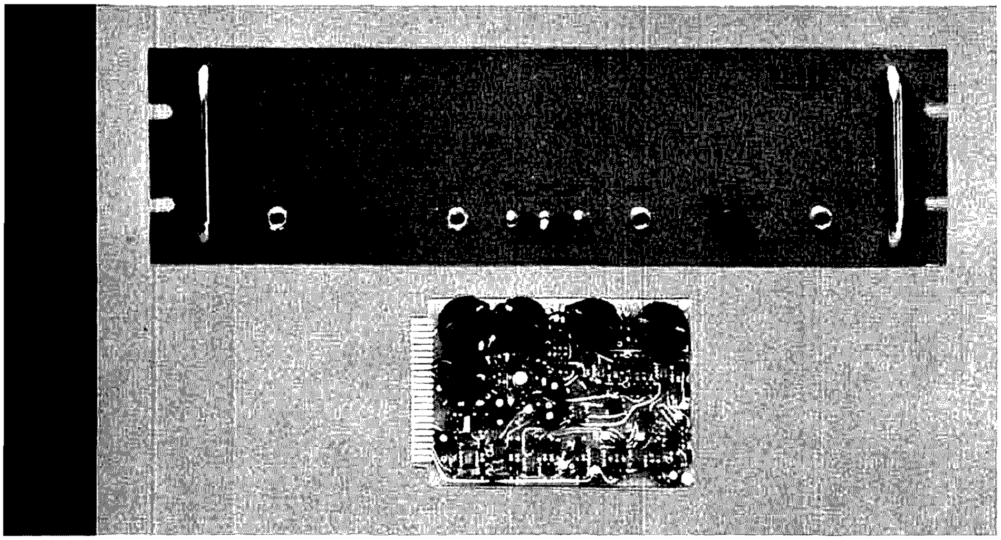
Docket Proposing that volunteer examiners be required to submit photocopies of their licenses with request for examinations has been released. Docket 20679, released December 22, was proposed to help establish the qualifications of volunteer examiners. Due date for Comments is February 2 and for Reply Comments February 12.

Applicants For An Extra Class License will no longer be required to have at least a year as a General or higher class licensee as a result of a Report adopted by the Commissioners this week. Effective date of the change was not available at press time.

AUTOMOTIVE IGNITION NOISE limits are being studied by FCC in an inquiry released by the Commissioners December 10. Docket 20654 relates results of an FCC-funded Stanford Research Institute study on low cost techniques for reducing impulse noise radiation from automobiles (FCC Report RS75-03, available from National Technical Information Service, Springfield, Virginia — order number PS239-471).

Comments On Effects of ignition noise on communications and other radio services as well as feasibility of radiation reduction are being sought by the Commission. Comment due date for Docket 20654 is March 19, 1976; Reply Comments are due May 4.

INSURANCE PROTECTION for mobile rigs may shortly become much stickier as thefts from autos skyrocket. Since January 1 all auto insurance policies in Texas will have endorsement A927 attached to them which states: "The insurance does not apply to loss of or damage to any device or instrument or a combination of devices or instruments designed as a Citizens Band radio, two way mobile radio or telephone, including its accessories, equipment or antennas." Though this limitation as yet applies only to Texas, the Insurance Services Office is reported to be seriously considering introduction of a similar exclusion nationally.



DT-600 RTTY demodulator

An advanced RTTY TU design with the most-wanted features — all on a single PC board

Amateurs have had much experience using IC RTTY demodulators but have found that certain optional circuits are difficult to add. Also, a number of options previously considered important were found to be no longer necessary. The ideas of active RTTY enthusiasts have been included in the DT-600 demodulator design described here.

Amateur RTTY has made significant progress since 1956. Only 850-Hz shift had been allowed until 1956, when the FCC revised regulations to allow any shift up to 900 Hz. It was found that properly designed and adjusted narrow shift (170 Hz) systems were superior to those with wide shift (850 Hz) in terms of adjacent signal rejection, weak-signal detection, selective fading, and noise immunity. The small group that began using narrow shift has grown to the point that wide-shift fsk is seldom heard in the high-frequency amateur bands so it may be assumed to be nonexistent for practical purposes. For this reason it's no longer necessary to provide the option of both narrow and wide shift in an RTTY demodulator. Thus requirements are eased for a sophisticated demodulator, with a resulting decrease in size, cost, and construction time.

It is also no longer considered necessary to provide an option for receiving inverted shift, as standards for direction of shift are well established (fsk *space* below *mark*), and upside down keying is seldom encountered. However, as current operating practice has simplified certain aspects of the RTTY demodulator, it has complicated others. The problem most experimenters face today is

By Robert C. Clark, K9HVW, Garey K. Barrell, K4OAH and Archie C. Lamb, WB4KUR

Any of the authors may be reached c/o the following address: 930 Chestwood Avenue, Tallahassee, Florida 32303.

modifying the demodulator to interface with external equipment such as the SELCAL,[1,2] a regenerator,[3,4,5] a speed converter,[3,6] a video monitor, or even a computer. This problem has usually meant ending up with scorched boards with lifted foil and components dangling in the air to make modifications to the existing board. The DT-600 provides TTL compatible DATA (mark/space) and AUTOPRINT (print/nonprint) outputs as well as provisions for remote motor control relay.

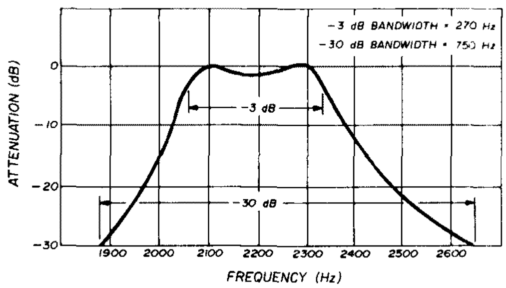


fig. 1. Bandpass filter response for narrow shift.

If you don't need the interface capability the interface components may be omitted, as the DT-600 will stand alone. If you wish, the high-voltage loop keyer, high-voltage loop supply, and motor relay may be mounted in the teleprinter and driven from the interface outputs. This option is particularly important when a number of machines at the operating location must be shifted from one loop to another. Keying and motor control can be easily controlled by a simple matrix switcher. Furthermore, if the high-voltage loop is completely contained within the teleprinter, there is much less likelihood of noise disturbing other equipment.

The DT-600 is an adaptation of the popular and reliable ST-6 demodulator,[7] incorporating the philosophy mentioned above. Other modifications have been made to reduce size, cost, and construction time of the DT-600. Additional design standards and philosophies of the CATC project, described below, have been incorporated. To meet the requirements of both amateur RTTY operators and the CATC project, the following features have been included in the DT-600:

1. Single-board construction

2. Single shift (may be either standard narrow 2125/2295 Hz or wide 2125/2975 Hz).

3. Optional interface connections.

4. Reduction of discrete components.

5. Choice of components to reduce size and cost.

6. A minimum of panel controls.

This article describes the DT-600, a single-board adaptation of the popular ST-6 demodulator introduced by Irv Hoff, W6FFC, in the January, 1971, issue of *ham radio*. The DT-600 will either stand alone or can be interfaced with a variety of other options. Its design results in a significant decrease in size, cost, and construction time with no decrease in performance. A future article will contain a brief description of the simplified but similar DT-500 vhf demodulator with examples of how to interface both units with external equipment. **editor**

For single shift, DT-600 performance is equivalent to that of the ST-6 but with the advantages described above.

The CATC group is charged with developing Computer-Automated Teletype Control for the Navy-Marine Corps MARS Teletypewriter Relay System. Its goal is to develop an automatic store and forward message system. The CATC group is using a systems approach for the development of receivers, demodulators, fsk and afsk generators, and control devices. The DT-600 is the first of a series of such devices to be described in coming months. It should be noted that designs described in this and future articles were by amateur radio operators (but professionals in electronics) on their own time. For this portion of the effort the CATC project provides only direction.

Much thought from CATC group members has been incorporated into the CATC system philosophy. Standards have been established for interface between equipment (TTL logic family compatibility), connections to card connectors, card configuration, and power supplies to allow for many options and ease of interconnection. The CATC philosophy allows the user to integrate system components (or self-designed equipment) into a working system with minimum cost, effort, and size, while allowing versatility, presently unavailable, for

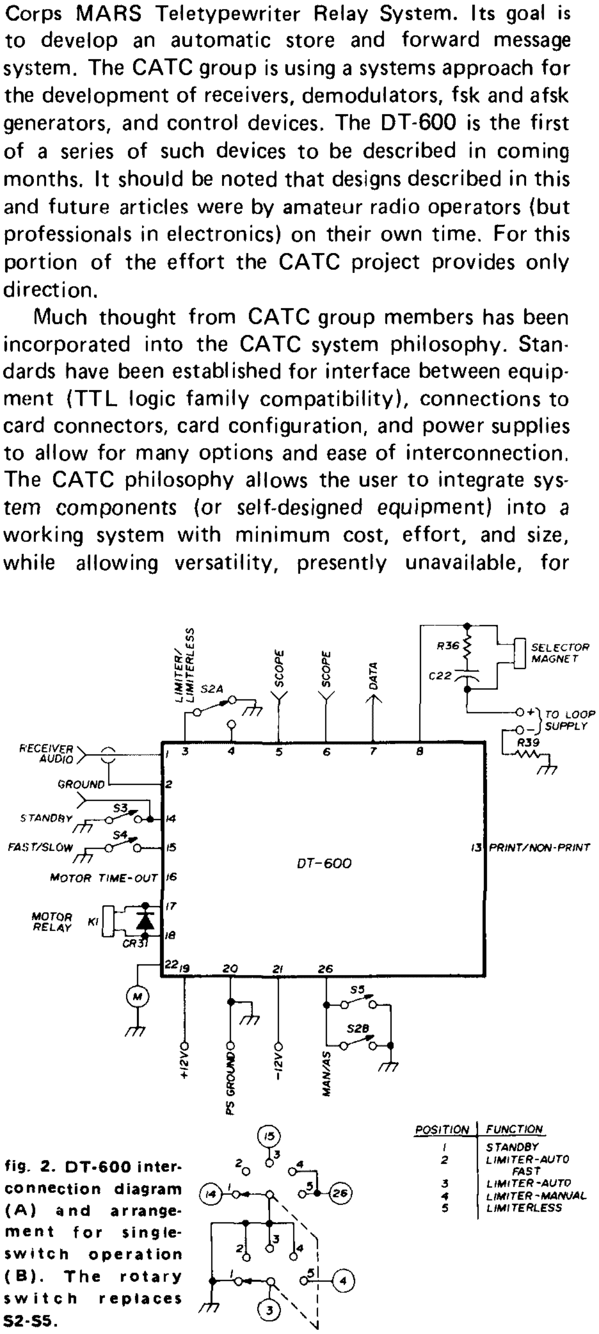
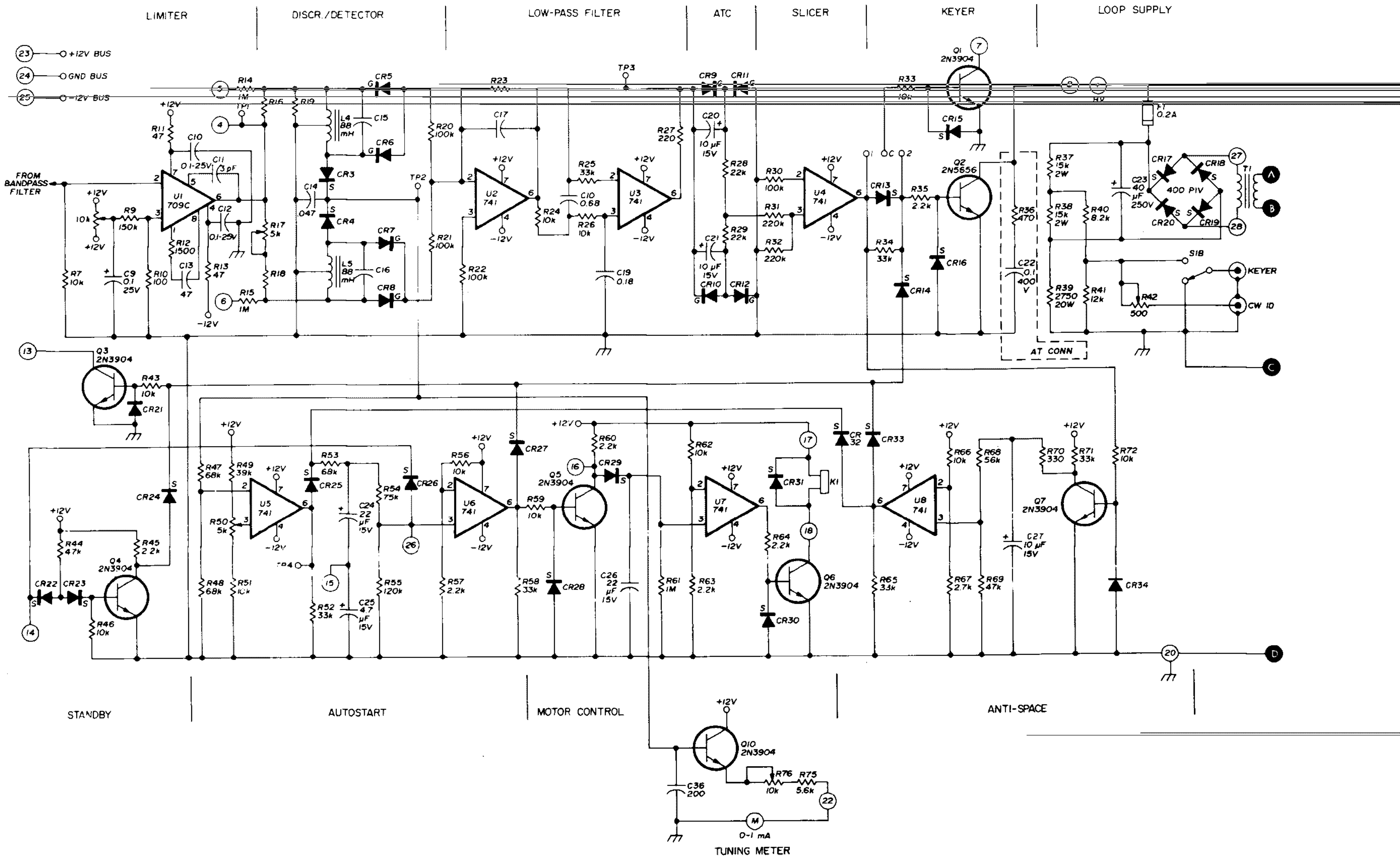
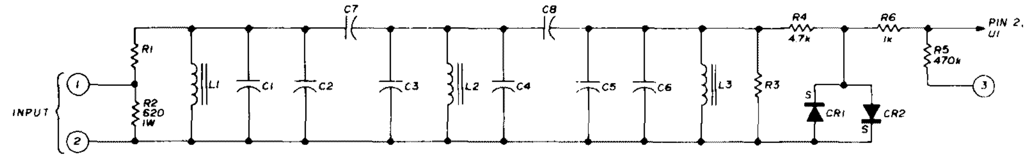


fig. 2. DT-600 interconnection diagram (A) and arrangement for single-switch operation (B). The rotary switch replaces S2-S5.


LIMITER
DISCR./DETECTOR
LOW-PASS FILTER
ATC
SLICER
KEYER
LOOP SUPPLY
+12V BUS
GND BUS
-12V BUS
FROM BANDPASS FILTER
AT CONN
KEYER
CW ID
STANDBY
AUTOSTART
MOTOR CONTROL
ANTI-SPACE
TUNING METER
0-1 mA



			bandpass filter values											tuning frequencies		
mark	space	shift	C1	C2	C3	C4	C5	C6	C7	C8	L1,2,3	R1	R3	sec 1	sec 2	sec 3
2125	2295	170	0.15	.056	.18		.15	.056	.022	.022	22 mH	1.6k	2.2k	2195	2195	2195
2125	2975	850	.015	.018	.015	.01	.015	.018	.015	.015	88 mH	2.7k	3.3k	2400	2500	2400

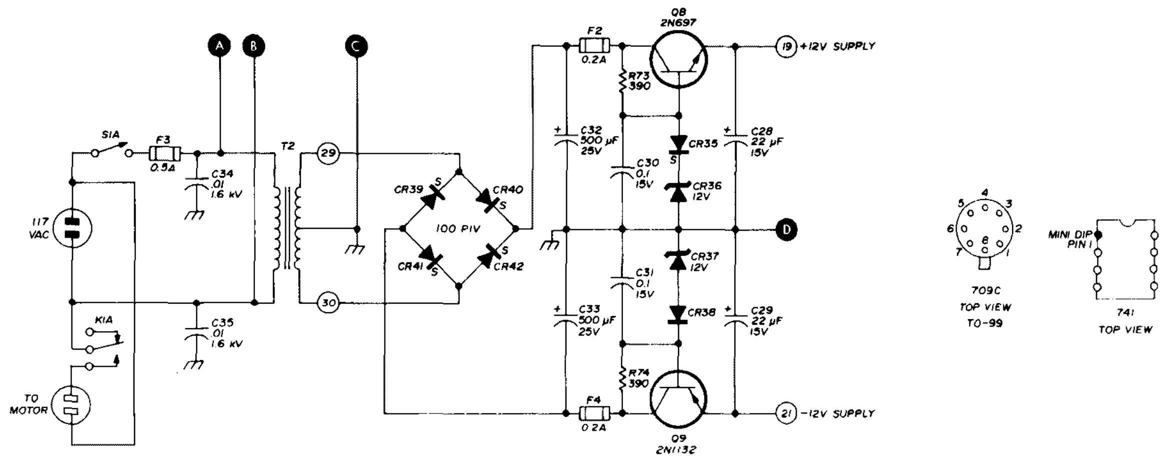


fig. 3. Schematic diagram of tne advanced DT-600 RTTY demodulator. All diodes marked G are Germanium 1N270; diodes marked S are silicon 50 PIV unless noted. All resistors are 1/4 watt. Transformer T1 is an Essex PA8421; T2 is a Triad F-40X or Essex P8180.

discriminator values

	2125/2295	2125/2975
R16	6.8k	4.7k
R18	6.8k	6.8k
R19	100k	33k
R23	270k	180k
C15	.068	.068
C16	.056	.033
C17	.012	.018

system expansion and modification. A simplified vhf demodulator (DT-500) will be described later.

The SELCOM (an advanced yet simplified multifunction version of the SELCAL) will be described in a future article. The SELCOM also functions as a regenerator and speed converter. A simplified mini-SELCOM, which provides the same features except for a limited number of functions, will also be described.

circuit description

The audio from the station receiver is introduced into the DT-600 through a three-pole Butterworth bandpass filter, **fig. 3**. This filter may be used for either wide shift (2125/2975 Hz) or narrow shift (2125/2295 Hz). The wide-shift bandpass filter is about 1 kHz wide at the -3 dB points, and the narrow-shift filter is about 270 Hz wide (**fig. 1**). This filter provides additional immunity to noise and adjacent signals even when the receiver has good selectivity. Also the bandpass filter significantly reduces any hum that may appear on the receiver audio, protects the first amplifier stage from being damaged by excessive audio input, and provides an impedance match between the receiver 600-ohm output and the high impedance of the first amplifier stage. CR1 and CR2 are ordinary silicon diodes that limit the audio at the junction of R4, R5, and R6 to 0.7 volt.

limiter

Audio from the bandpass filter is amplified by U1, a 709 operational amplifier in the open-loop configuration. U1 functions as an amplifier and hard limits the signal to ±10 volts (at the output). Output from pin 6 is a square wave as long as the input signal exceeds the extremely low limiting threshold (about 1 mV). Thus, very large changes in the rf or audio signal levels may be tolerated. R8 establishes the balance for U1 to provide for minimum threshold and symmetrical output. R12, C13, C11 provide frequency compensation, while R11, R13, C10, C12 decouple U1 from the dc supply lines. Limiterless operation is available by connecting point 3 to point 4. In this configuration U1 limits only on signal peaks. Note that autoprint operation would be unreli-

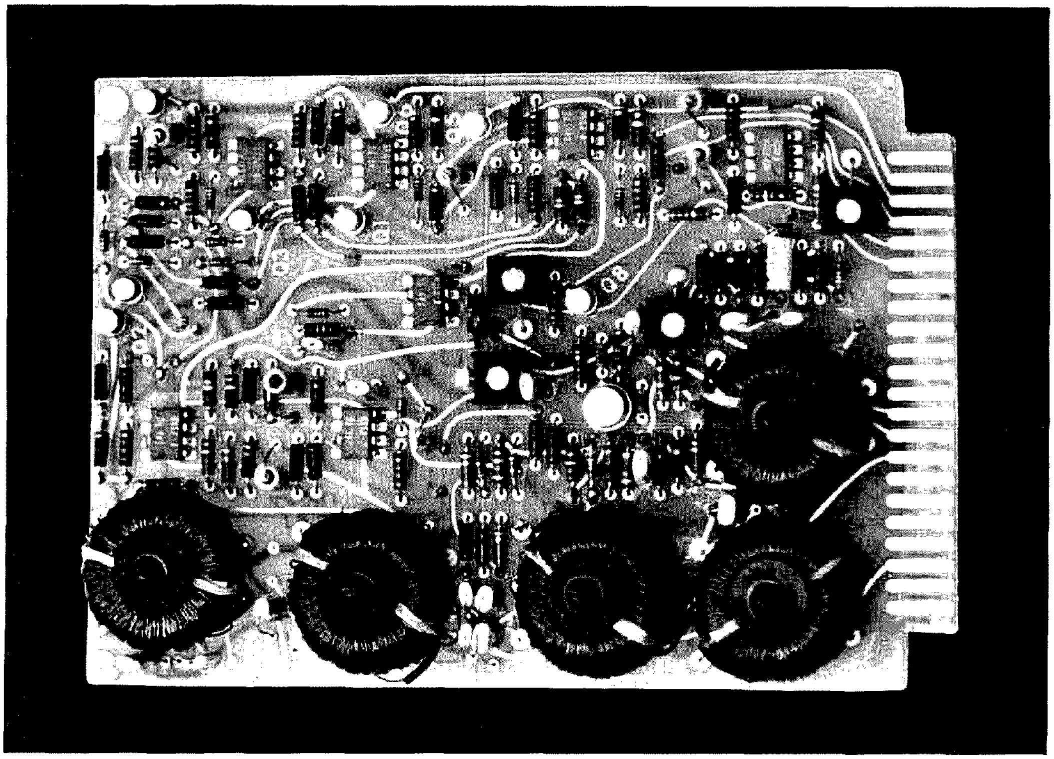
able in this mode, so the motor and print control sections are forced on (fig. 2) by a parallel switch section.

discriminator/detector

The square wave output of U1 is coupled to the discriminator via R16, R17, R18. R17 is set to equalize the voltage levels of both mark and space signals in the discriminator. L4, L5, C15, C16 form a linear discriminator. Its narrow-shift response is shown in fig. 4, with peaks about 100 Hz wide at the −3 dB points. L4, C15 form a parallel-tuned circuit at 2125 Hz (mark). L5, C16 are tuned for the space frequency (2295 Hz for narrow shift or 2975 Hz for wide shift). Since the same inductance is used for both mark and space frequencies, the filter Q would be different, which in turn would cause unequal bandwidth for the two filters (equal bandwidth is particularly important for limiterless copy). R19 damps the Q of the L4, C15 combination to match the Q and bandwidth of the L5, C16 tuned circuit. CR5, CR6, CR7 and CR8 (1N270 germanium) form full-wave detectors for minimum ripple. CR3 and CR4 form an OR gate so that a positive voltage appears across C14 if either mark or space tones are present. This voltage is used to control the demodulator autoprint section and may also be used to drive a tuning meter. Scope connections are through one-megohm resistors to eliminate loading the discriminator.

low-pass filter

The two-stage (741 op amps) active low-pass filter is designed for 100 wpm operation. Degradation of 60 wpm performance is so slight with the filter set for 100 wpm (fig. 5) that the additional complexity of switching filter characteristics to optimize response for 60 wpm is not considered worthwhile. If only 60 wpm is to be used, R24, R26 may be changed to 16k, and C17 to 0.02 μF for narrow shift and 0.03 μF for wide shift.



Complete DT-600 RTTY demodulator is built on one plug-in printed-circuit board. The five 88-mH toroids used in the filter circuits are mounted along the lower edge of the board. Double-sided circuit boards with plated-through holes are available from Data Technology Associates (see footnote on page 14).

ATC and slicer

During selective fading the automatic threshold corrector provides the symmetry necessary for the slicer, another 741 op amp. Action is shown in fig. 6 for selective fading on the space channel. Without the ATC action selective fading would bias the signal. The symmetrical ATC output is fed to U4 which is set for maximum gain to provide uniform keying from the varying, but symmetrical, ATC output. This high gain allows just 1 or 2 mV over offset to trigger the output and allows copy during deep selective fading and incorrect (straddle-tuned) shift.

keyer

Strapping options are available on the board so that slicer output may be fed directly to the base of Q1 (jumper from 1 to C), an open-collector output stage.

Output from Q1 does not respond to the mark hold provided by the standby line, autoprint section, or anti-space. Q1's output may be used to provide data to external equipment, such as the SELCOM, where the autoprint attack time delay is not desired. CR15 prevents the negative portion of U4's output from reaching the transistor.

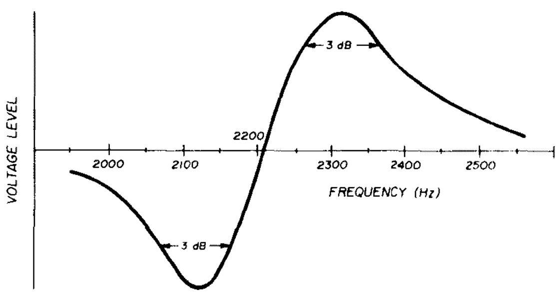


fig. 4. Discriminator response for narrow shift.

By strapping 2 to C, Q1 may be inhibited by the standby line, the autoprint section, and the anti-space section. Q2 is the high-voltage keyer. It is always inhibited, as described for the second option on Q1. For a simple selector magnet keyer stage, the importance of a high-voltage supply and keyer can't be over emphasized.[8] Several designs are based on the fact that only 12 volts are necessary to maintain a 60-mA selector-magnet current but completely overlooked is the behavior of a large inductance (selector magnet) in an ac (switched) circuit. Results from such low-voltage keyers are very poor, even with no distortion on the received signal. Their operation deteriorates rapidly with distortion. Q2 may be omitted and Q1 used to drive a remote keyer within the teleprinter, as mentioned earlier. (A simple circuit to control the loop keyer and motor relay through logic levels will be described in the future article which features the simplified DT-500 demodulator.)

R36 and C22 suppress keying transients, and CR16 prevents the negative-going pulses from the selector-magnet field decay from being propagated back through preceding stages. It's possible through the options provided on the board to take the data signal from Q1, process it in external equipment, then reintroduce the processed signal to the loop keyer. (This feature will be used with the DU-200 regenerator and speed converter to be described as the basis for the SELCOM in a future article.)

autoprint

CR3 and CR4 form an OR gate as mentioned earlier. If either mark or space tones are present, a positive voltage will appear across C14. U5 threshold is set (by R50) so that with no signal present its output is positive. This positive voltage locks the keyer stage (through CR27) in mark hold. When a signal appears on either mark or space frequencies, the OR gate output forces U5 to reverse state so that its output is negative. This action stops the charge on C24, C25, which begin to discharge through R54, R55. When the voltage at point 26 has decreased to a level determined by R56, R57 (time constant is such that it takes about 1.3 seconds to reach this level with C24, C25 in series and about 7.4 seconds with point 15 shorted to ground), U6 pin 6 is forced negative, the standby line is released; Q5 is biased off charging C26; U7 output is forced positive; and Q6 conducts, pulling in motor relay K1. If the signal disappears, or if (as in CW) the duty cycle drops below 25%, then C24, C25 begin to charge, eventually returning to mark hold. However, C26 must discharge below the voltage on U7 pin 2 before the motor relay is released. This delay gives about 38 seconds after loss of signal before motor shut down, which is desirable to keep the motor from turning on and off between transmissions or when the signal fades into the noise. Shorter turn-on and turn-off times may be had by making C24 and C26 each 18 μF.

anti-space

In the mark condition, the positive output of U4 (pin 6) forward biases Q7, preventing C27 from charging. On space, Q7 is shut off and C27 begins to charge. The time constant is such that C27 will charge above the threshold set for U8 in slightly over 132 ms (132 ms is the longest steady space expected from valid RTTY, a blank at 60 wpm). Thus, U8 will not trigger with normal RTTY where C27 is quickly discharged by each mark signal through Q7. Should a space exceed the time constant, U8 output becomes positive, and the positive voltage is applied through CR33 to the mark hold line. The positive output of U8 is also applied through CR32 to C24, C25, starting the motor time-out sequence. Thus a steady space will a) not turn the machine on, b) immediately clamp the printer in mark hold if it is already on, and c) begin the time-out sequence.

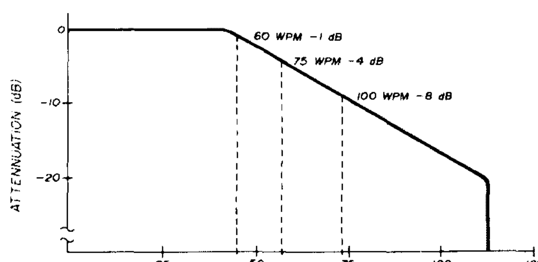
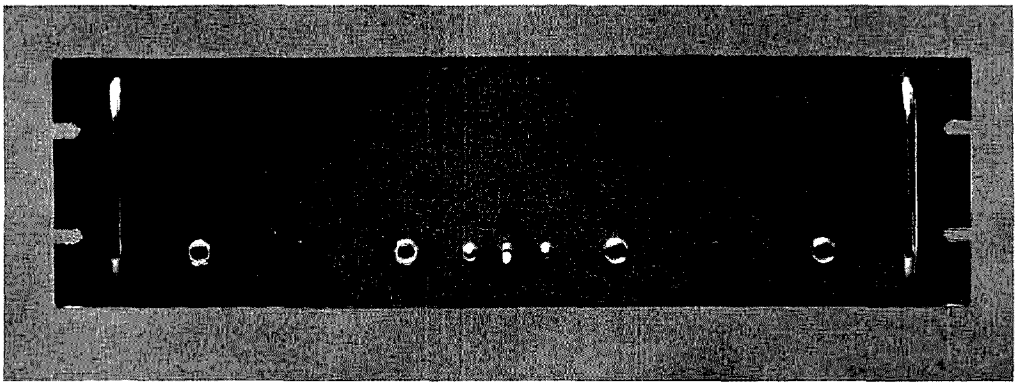


fig. 5. 74.18 baud low-pass filter response.

standby

When the standby line (14) is grounded, Q4 forward bias is removed, placing a positive voltage on the mark hold line through CR24. Also U6 pin 3 is pulled low, starting the motor, as described previously. (The only time the standby feature is regularly used is during trans-

mit and, hence, holds the motor on during a long transmission.) Note that the mark hold line is positive when no signal is present, the standby line is grounded, a steady space is present, or a signal is present (CW or other non-RTTY) that doesn't switch U5. All these situations occur on nonprint. The mark hold line positive voltage forward biases Q3, which may be used to indicate to external equipment that no valid RTTY signal is present. When a valid RTTY signal does appear, Q3 forward bias is removed.



The DT-600 RTTY demodulator described in this article is only part of the overall CATC system of RTTY modules. Others will be described in future articles.

tuning meter

The CR3, CR4 OR gate output is also coupled to the base of Q10, a meter amplifier. When the RTTY signal is properly tuned, a positive bias is applied to Q10 base if the signal is either mark or space. The tuning meter reading will be proportional to the signal level of mark (or space) at the discriminator. In operation the receiver is tuned for the highest steady meter reading. For tuning incorrect shift the meter tuning indicator is superior to an oscilloscope, as a proper meter reading will closely indicate balanced output from the discriminator to the low-pass filter.

loop power supply

A full-wave 170-volt supply provides the required loop current (60 mA). Note that loop-current limiting resistor R39 is in the negative supply lead. This is the floating loop introduced by Hoff and included in the ST-6.[7] By allowing the supply to float, a polar output (mark negative, space positive) is available at point 11 to key either an afsk or fsk oscillator. Grounding point 12 gives less than full saturation current through a shifter diode, yielding narrow-shift CW identification.

components

The DT-600 is constructed on a single 4½ by 6 inch (11 by 15cm) board, which includes all parts except the switches and power supplies (a single ±12 volt supply may power several DT-600s).* A 22-pin edge connector is provided at the board edge.

The capacitors that tune the bandpass and discriminator filters should be of high quality and have high Q. Sprague *Orange Drops* or Mallory polystyrenes are generally available and recommended. Tantalum capacitors are recommended for C20, C21, C24, C26, and C27. All resistors are ¼ watt unless otherwise indicated. Diodes CR5 to CR12 are germanium 1N270s. All other diodes may be 1N914 or equivalent (note: larger diodes will not fit the available space on the circuit board). CR17 to CR20 are 400 PIV 1 amp, and C39 to C41 are 100 PIV, 1 amp. U1 is a 709 op amp (must be in TO-99 package), but all the others are 741s (8 pin mini-dip). Substitutions should not be made. Transistors are specified, but only Q2 is critical. Nearly any transistor that meets the specifications will do (2N3904: npn silicon switch V_{ce} = 40 V at 200 mA; 2N5656: npn silicon switch V_{ce} = 300 V at 500 mA). The loop supply transformer is an Essex PA-8421, which provides 125 Vac at 50 mA, but since the filament winding isn't used the transformer is well within rating, supplying 60 mA to the loop.

Relay K1 should have a 12-volt coil and the contacts should be rated at 10 amps for long life, such as the P&B KA11DG. Trimmers R8, R17, R50, R76 are PC-board mounts, such as TRW X201. A single two-pole, five-position rotary switch can handle all the switching functions (fig. 3B), or miniature toggle switches may be used. M1 is any inexpensive 0-1 mA meter. Dale EBT156 22-pin connectors may be used for the edge of the card.

construction

Consult the parts layout sheet provided with the circuit board and mount all parts on the board except for components for the bandpass input filter and discriminator. These tuned circuits must be adjusted to the proper frequency before they are permanently mounted on the board. The tuned circuits are adjusted with an audio

*A double-sided printed-circuit board with plated-through holes for the DT-600 is available from Data Technology Associates, Inc., Post Office Box 1912, Miami, Florida 33143. The price is $12.50, postpaid.

oscillator and a frequency counter coupled to the tuned circuit through a high-resistance (100k) to eliminate loading of the tuned circuit (fig. 7). Audio voltage across the tuned circuit is monitored by a vtvm. The audio generator is adjusted to obtain the peak. If the frequency is lower than desired, reduce the inductance to increase resonant frequency. If the frequency is higher than desired, capacitance may be added. Note that the inductors for the narrow-shift bandpass filter are 22 mH. These may be formed by placing the two windings of an 88-mH toroid in parallel.

filter tuning

The procedure for tuning the narrow-shift bandpass filter is to mount all capacitors on the board, but omit R1, R2, R3, and R4 at this time. Each of the three sections is, in turn, tuned to 2195 Hz. Short the toroid in the center section and tune the first and third sections to the desired frequency. Then remove the short from the center section, short the toroids in the first and third sections, and tune the center section to the desired frequency. Remove all shorts and place R1, R2, R3, R4 in their respective positions on the board. The discriminator filters are best tuned on the board, supplying either mark or space tones through the input filter. Use care in tuning all filters, as performance will be seriously degraded if filters are not resonant at the proper frequencies. No instructions are provided for tuning the wide-shift input filter, as it is sufficiently noncritical with the specified capacitors that tuning is not required.

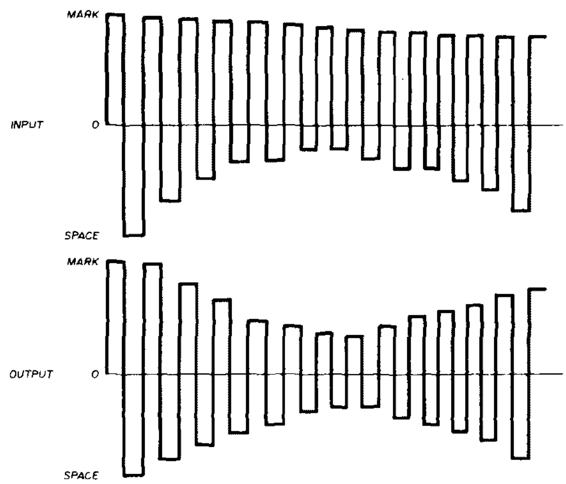


fig. 6. Automatic threshold corrector action for selective fading on space.

adjustments

After the board has been completed and filters tuned, the unit may be adjusted. Short the audio input and adjust R9 until the voltage at TP1 is zero. Remove the short from the audio input and apply a mark tone. Note the reading at TP2. Adjust the audio oscillator for the space frequency (depends on choice of shift), and adjust R17 for the same reading at TP2 as obtained with the mark signal. Repeat the procedure until the readings are identical for both mark and space. R50 determines the bandpass for autoprint (i.e., how far off frequency a station can be and still hold in the autoprint). Set the audio oscillator about 50 Hz below the mark frequency and adjust R50 so that the voltage at TP4 fluctuates near zero (both positive and negative excursions). R76 should be set for a maximum meter reading of 70% of full scale for either mark or space.

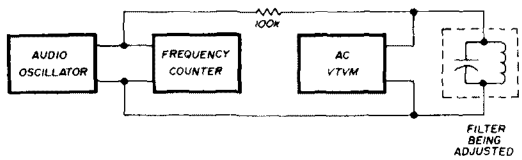


fig. 7. Instrumentation for filter tuning.

With all adjustments completed you are now ready to operate the DT-600. In normal operation nothing is done to the DT-600 except to ground the standby line during transmit. For very weak signals you may prefer to switch to limiterless and manual print control. Sometimes it's a bit frustrating to think that there's no way to adjust things to improve the print. However, with the exception noted above, you can be sure the DT-600 is providing the best print available for the price, and only slight improvement is available even at the higher prices. Interface connections will be discussed in a future article.

acknowledgements

We wish to thank Werner Fehlauer, KL7HKB, for constructive comment on early designs; Ronald C. Viets for parts layout and PC artwork changes; James E. Scalf, K4TKU, for drawings (and their many revisions); Fred R. Scalf, Jr., K4EID, for CATC systems interface compatibility and project coordination.

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ham radio

a new look at solid-state amplifiers

Techniques for joining unipolar and bipolar transistors to exploit the features of each

This article reveals no technological breakthroughs, nor will it lead you through the intricacies of a construction project. Rather, it suggests that amateurs and experimenters have overlooked a useful and versatile cricuit technique – the marrying of unipolar and bipolar transistors to produce an amplifying module with the desirable features of each solid-state device. In the following paragraphs arguments in behalf of the union of these devices are developed. I hope these discussions will stimulate the interest of those who enjoy designing and building their own equipment.

When bipolar transistors first became commercially available, it became obvious that this device had a serious shortcoming compared with the vacuum tube: the current hungry base-emitter junction was recognized as a sorry trade for the voltage-actuated input circuit of the tube. Because of the many advantages of the transistor (no heater, no microphonics, negligible aging, small size and cost) we learned to live with its low-impedance, power-consuming input.

fet transistor

Technological evolution produced the unipolar, or field-effect transistor, known as the fet. Logic can be presented to show that the fet should have chased the bipolar transistor right off the market. Some of the reasons why such a displacement did not occur are:

1. Fets tended to lag behind bipolars in gain-bandwidth product.

2. Fets acquired a reputation for being limited in power-handling capability, even for the needs of low-level circuitry.

3. Fets have never been cost competitive with bipolar transistors.

4. At least until recently, fets have not been hot performers – transconductance tended to be low – in the several hundred to several thousand micromho range.

5. The electronics fraternity has been in need of articles such as this to illustrate the benefits of beefing up fet performance with the bipolar transistor.

operational amplifier

What about the operational amplifier? Surely, the monolithic op amp *must* be the ultimate amplifying device. Not necessarily! From the viewpoint of the experimenter, the op amp has the following disadvantages:

1. It is far from easy to work with unless your eyes, nerves, and fingers were predestined for the jewelry trade.

2. During experimentation, it is vulnerable to catastrophic damage.

By Irving M. Gottlieb, W6HDM, 931 Olive, Menlo Park, California 94025

3. Dual-polarity dc supplies are required.

4. The cheapies — the ones amateurs can afford — are notorious for performance kinks such as latch-up from overdrive and a propensity for oscillation.

5. The op amp is a bargain, it's true, in terms of the perhaps several-dozen discrete devices displaced. But a great number of amplifying tasks don't require differential input, dc response, or accurate operational functions. As a more mundane gain-producing device the op amp often is less than a good buy.

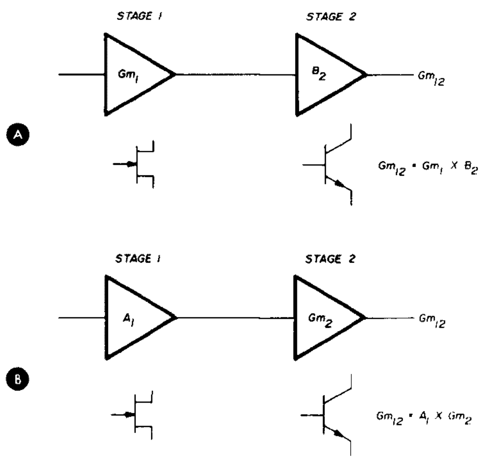


fig. 1. Transconductance amplification in a two-stage amplifier. G_m is increased by the current gain of a following stage, A, and by the voltage gain of a preceding stage, B.

We could deal similarly with the merits and shortcomings of other amplifiers; for example the tube, magnetic amplifier, and tunnel diode. All have problems for general experimental use. An amplifying module with the high-impedance input of tubes and the output characteristics of bipolar transistors would be a major step in the right direction.

the transconductance problem

One of the salient features of the bipolar -- and one not generally appreciated — is its inordinately high transconductance, which can range from several hundred thousand to millions of micromhos. Think of a tube or fet with such a characteristic! The reason that little awareness of this feature exists is that the healthy transconductance loses much of its significance when we have to supply power to the input circuit. Now the concept of driving a bipolar with an fet should begin to make sense. With such a combination we can achieve both high input impedance and high transconductance.

In the amplifier cascade of **fig. 1A** assume that the transconductance of stage 1, the fet, is Gm_1, and that the current gain of stage 2, the bipolar, is B_2. The *overall* transconductance of the cascade is given by $Gm_1 \times B_2$. Expressed in words, the transconductance of a stage is increased by the current gain of a *subsequent* stage. (Keep in mind the concept of transconductance as the figure of merit of amplifying capability.)

Let's now deal with the amplifying cascade depicted in **fig. 1B**. This time, the voltage gain of stage 1 is known and is represented by A_1. The transconductance of stage 2 is represented by Gm_2. The overall transconductance of the amplifying cascade is given by $A_1 \times Gm_2$. In words, the transconductance of a stage is increased by the voltage gain of a *preceding* stage. Assuming that the same amplifying cascade is represented by the block diagrams of A and **B** in **fig. 1**, some meaningful insights can now be attained.

We have seen that two products are both equal to a common quantity: overall transconductance, or Gm_{12}. We can therefore write:

$$Gm_1 \times B_2 = Gm_{12} = A_1 \times Gm_2 \tag{1}$$

or simply

$$Gm_1 \times B_2 = A_1 \times Gm_2 \tag{2}$$

We can make any of four algebraic transpositions; that is, **eq. 1** can be manipulated to facilitate the solution of any of its four terms. For our investigations, a particularly interesting transposition focuses on Gm_2, the transconductance of the bipolar transistor. Thus, we have:

$$Gm_2 = \frac{Gm_1 \times B_2}{A_1} \tag{3}$$

practical example using a fet and a bipolar

When we consult manufacturer's spec sheets, we generally find the transconductance of fets and the current gain of bipolars. (We seldom find the transconductance of bipolars.) Let's deal with the type 2N5438 n-channel fet and the 2N3565 npn bipolar transistor. The fet can have a nominal transconductance of 4000 micromhos, and the bipolar can have a nominal current gain of, say, 300. (The parameter tolerances of both devices are, to say the least, sloppy.) Suppose that the fet is used as a common-source voltage amplifier and that its voltage gain is four. It is thus employed to drive the bipolar. With a bit of crank-grinding, we can use these numbers to compute Gm_2, thus:

$$Gm_2 = \frac{4000 \times 300}{4} = 300{,}000 \text{ micromhos}$$

transconductance for the 2N3565 transistor

Note, too, that the overall transconductance, Gm_{12}, of the amplifier cascade calculates to a whopping 1,200,000 micromhos! (Both $Gm_1 \times A_1$, and $A_1 \times Gm_2$ confirm this result.) Whether a single device or a module with more than one active device, an amplifier that develops over a million micromhos and extracts no power from the signal source has to be what the doctor ordered. Also to be considered is the fact that voltage gain is available from the bipolar. If we insert a 1k load resistor in the bipolar transistor collector circuit, the circuit will develop a voltage gain of 300 (from voltage gain $= Gm \times R_L = 0.3 \times 1000 = 300$, where Gm is expressed

in mhos). The overall voltage gain of the amplifier cascade is then the product of the voltage gains of stages one and two, or 4 X 300 = 1200. This is confirmed by multiplying the overall transconductance, Gm_{12}, by the output load resistance, or 1.2 X 1000 = 1200. This calculation is made on the premise that a 1,200,000 micromho, or 1.2 mho, amplifier acts upon a 1000-ohm output load resistance. Note that high voltage gain can be produced in the bipolar stage with low load resistances, which implies relatively low degradation of higher frequencies. If you wanted high voltage gain in a single fet, the drain resistor would have to be many tens, perhaps hundreds, of kilohms; and frequency response would peter out pronto.

multipurpose amplifying module

In **fig. 2** we have an amplifier with the described performance characteristics. The −3 dB points are approximately 100 Hz and 0.6 MHz. But this is only a start. The circuit is extremely flexible. The frequency response, voltage gain, power output, and power consumption are easy candidates for selective optimization. Such versatility and noncritical features stem from the use of ac coupling between the stages. Direct coupling can also be used but will, in general, require a bit more patience in satisfying the mutual bias requirements of the two active devices. Direct coupling can lead to more compact packages and is necessary, of course, if dc amplification is needed.

The bipolar load resistor, R_L, is in effect acted upon by an overall transconductance exceeding a million micromhos. At the same time, the input impedance is of the same order of magnitude obtainable from vacuum-tube amplifiers and is limited only by R_G. The dashed enclosure in **fig. 2** facilitates thinking of the cascade as a *single* "amplifying module." Component values are noncritical and can be modified considerably from those indicated to optimize gain, frequency response, power output, or power consumption. Similarly, other than the indicated devices can be used. Not obvious from inspection of **fig. 2** is the fact that the load presented to the fet is primarily the input resistance of the bipolar transistor. This value is in the order of 1000 to 1200 ohms and enables the fet to develop a voltage gain of 4.

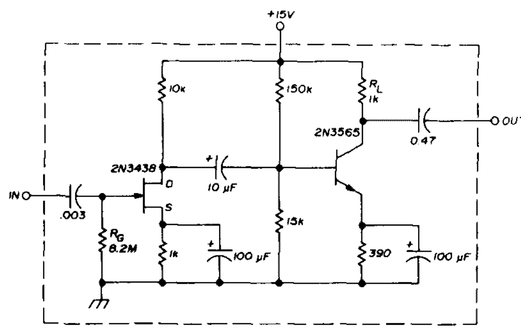


fig. 2. Multipurpose amplifying module using a field-effect transistor to drive a bipolar transistor.

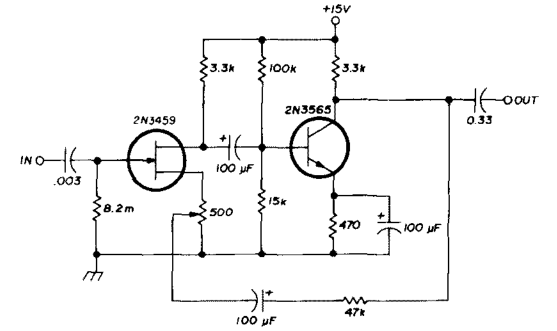


fig. 3. Low-distortion feedback amplifier. Circuit can be optimized for rf as well as audio frequencies by adjusting coupling, feedback, and emitter bypass capacitor values.

feedback amplifier

In **fig. 3** a similar amplifier is shown, but with the addition of an overall feedback path. Depending on how much feedback is used (how much the overall gain is decreased), various circuit attributes are evident. These include distortion reduction, extension of frequency response, and stabilization against gain variations, which otherwise tend to occur from the effects of temperature on both active and passive components and from changes in power-supply voltage. In this circuit, the feedback decreases the amplifier output impedance, which is usually a desirable feature.

This amplifier, like the previous one and the subsequent ones as well, can be optimized for rf as well as audio. In this particular case, you would reduce the size of the coupling, feedback, and emitter bypass capacitors. Or if both low and high frequency response are desired, these capacitors can be paralleled by small ceramic or mylars (electrolytics often don't perform well at higher frequencies). At high frequencies, the amplifier physical layout becomes exceedingly important, and a printed circuit board is probably the best approach.

Other things being equal, the extent of flat frequency response increases with increased feedback. If in addition to these techniques the fet and the bipolar are selected for both high transconductance and gain-bandwidth product, such a feedback amplifier can provide voltage gain by a factor of several tens to a frequency of several or more MHz. (For higher frequency work, the cascade arrangement of **fig. 4D** is best.)

experimental amplifying-module family

Nine other unipolar-bipolar amplifying modules are shown in **fig. 4**. Their names and applications are:

A. Alternative feedback amplifier – audio, general purpose, rf capability at low gain.

B. Audio amplifier with direct coupling – speech amplifier, low-level audio.

C. Audio amplifier for operation from rectified line voltage – audio output.

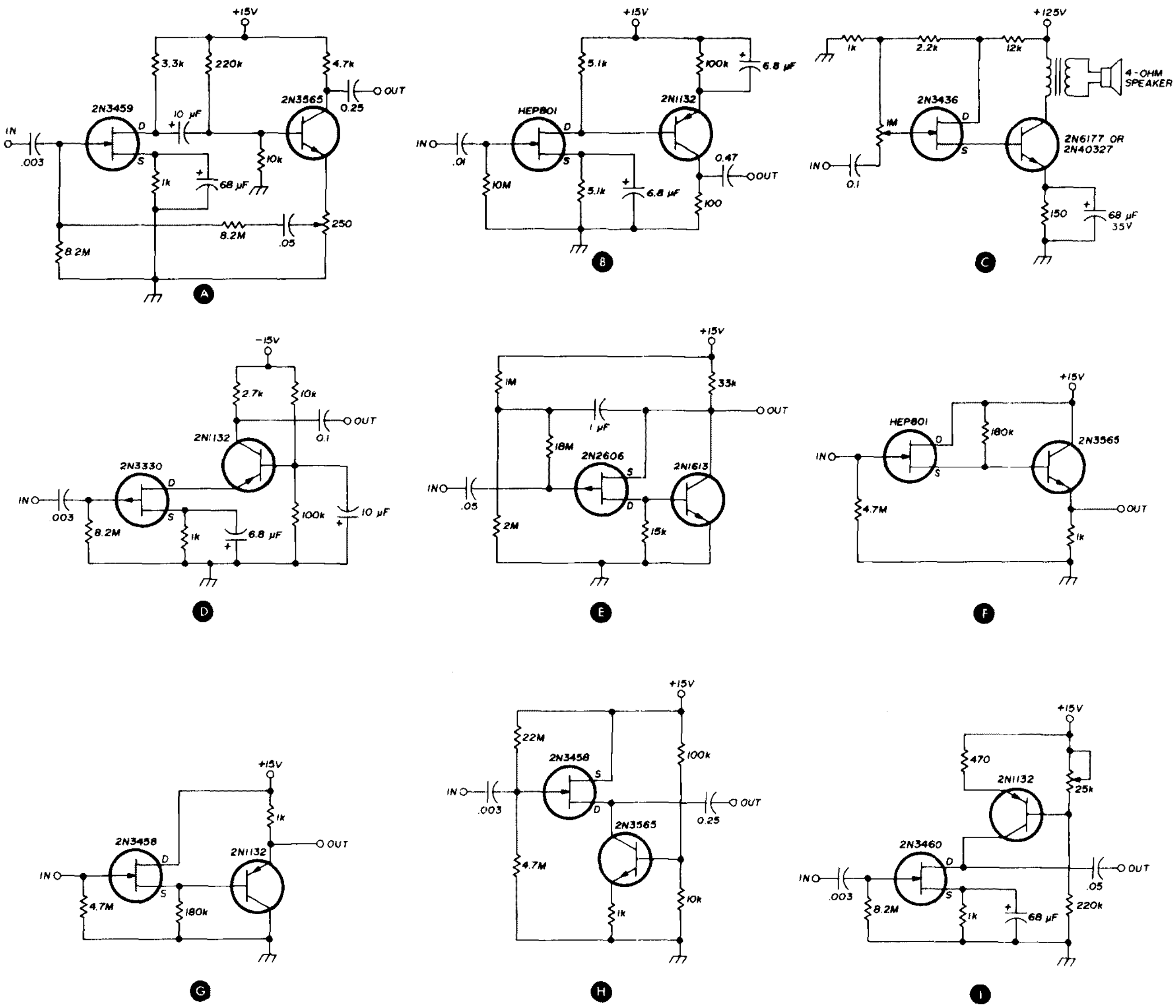


fig. 4. Family of amplifying modules useful for many applications.

D. Cascode amplifier – audio, video, i-f, rf; best arrangement for use with tuned circuits.

E. Ultrahigh input impedance amplifier – active scope probe, electrometer, instrumentation.

F. Darlington amplifier – meter interface, impedance transformer, coax driver, relay actuator.

G. Complementary symmetry Darlington – meter interface, impedance transformer, coax driver.

H. Source follower with constant-current bias supply – used where a source follower with high output-voltage swing and voltage gain close to unity is required.

I. Single-stage amplifier with dynamic load – high voltage gain; can be used with very low supply voltage.

In all instances the designated device types and component values are intended only as a guide. Because of device and component tolerances as well as the specific performance required, various modifications will probably be made. In particular, the empirical determination of bias networks in direct-coupled circuits will usually pay dividends in the attainment of symmetrical voltage swing. Improved performance of these circuits, as well as those in **figs. 2** and **3**, can sometimes be obtained by connecting a high resistance from the fet gate to the ungrounded battery terminal. Several tens of megohms should do the trick.

Why not build a few of these amplifier modules and retain them as convenient building blocks?

ham radio

vestigial sideband microtransmitter for amateur television

Amateur television video bandwidth can be reduced by adapting commercial techniques

To conserve spectrum space commercial television uses a transmission mode known as vestigial sideband. A composite video signal, containing frequency components from dc to 4 MHz, is amplitude modulated onto a carrier. The resulting sum and difference frequencies (sidebands) occupy an 8-MHz bandwidth. Before transmission, the modulated signal is filtered. The upper sideband and carrier are transmitted, but most of the lower sideband is not (see fig. 1). Thus the video signal plus its audio can be transmitted in the 6-MHz TV-channel allocation.

As amateur television (ATV) activity expands in the 70-, 23- and 13-cm bands, it will become necessary for amateurs to adopt vestigial sideband as their operating mode to avoid interference with other communications services. A case in point is the possibility of interference with the 435.1-MHz OSCAR satellite telemetry beacon, which would result from the unfiltered lower sideband of an ATV station operating on the 439.25 MHz ATV calling frequency.

In commercial television, the modulated carrier is developed, and filtering performed, at the ultimate transmission frequency (fig. 2). A complicating factor, the need for frequency flexibility, makes such a system impractical for ATV. Imagine retuning a stagger-tuned string of over-coupled resonator pairs for sharp skirts and flat response over a 5-MHz band, then retuning it each time you need to shift your operating frequency!

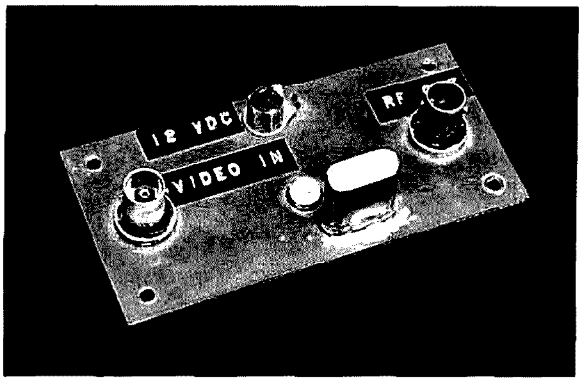
One alternative is to generate a stable, well-filtered vestigial sideband video signal on a fixed frequency in the vhf spectrum, then heterodyne it to the desired uhf in a balanced mixer. The conversion stage local-oscillator chain, if made variable in frequency, will provide the system with the required frequency flexibility. **Fig. 3** is a block diagram of one such system, which I use for ATV transmission in the 70-cm band. The observant reader may note in fig. 3 a pronounced similarity to the transceive converter for 1296-MHz ssb published in an earlier issue.[1] Obviously, the process of heterodyning a modulated signal into a higher frequency band for transmission is virtually the same, regardless of whether the original signal was modulated with a-m, fm, ssb, CW, or video.

By *H. Paul Shuch, WA6UAM,* *Microcomm, 14908* Sandy Lane, San Jose, California 95124

Many of the blocks in the local oscillator and rf strings of fig. 3, as well as the mixer, are either available commercially or may be adapted from equipment designs published previously. This article deals with the design and construction of the microtransmitter and vestigial sideband filter modules of the ATV system in fig. 3 — building blocks toward clean, commercial-quality TV transmission.

microtransmitter chip

The heart of the ATV transmitter is the LP-2000, a miraculous integrated circuit from Lithic Systems Inc., in Saratoga, California.* The outgrowth of a program to develop a microminiature aircraft crash-beacon transmitter, the LP-2000 is a complete transmitter system — oscillator, buffer, driver, power amplifier, modulator, preamplifier and regulator — all in a single 10-lead, TO-100 can. With the addition of a crystal, two tuned circuits, a battery, and a modulation source these ICs can generate as much as 100 mW of CW, or 50 mW of a-m or pulse-modulated output well into the vhf spectrum. **Figs. 4 and 5** indicate the very complex circuitry that can be built into a single monolithic microcircuit. A complete



Heart of the ATV system is the LP-2000 IC next to the crystal.

circuit description is available from the manufacturer in the form of an application note.[2]

An appealing feature of the LP-2000 is that its modulator transistors (Q14 and Q16 in fig. 5) are dc-coupled to both the driver and power amplifier transistors, Q13 and Q15. Additionally, direct coupling is employed between all modulator stages. Thus the circuit lends itself well to video-modulated applications.

frequency selection

The operating frequency chosen for the microtransmitter, 61.25 MHz, corresponds to the assigned video carrier frequency of commercial TV channel 3. This permits the basic microtransmitter module to be used for short-range, closed-circuit TV applications, there being no local channel 3 allocation in my area to interfere with such operation. Similarly, you may wish to select an operating frequency corresponding to the video carrier

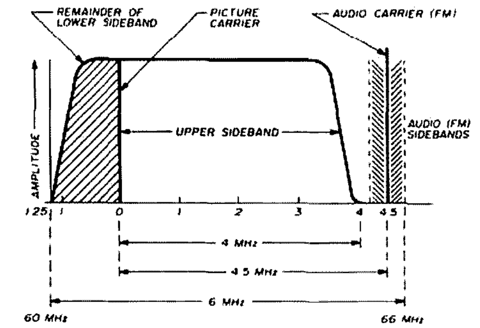


fig. 1. Vestigial sideband transmission on monochrome television signals.

frequency for a locally unassigned lower vhf-band TV channel.

The circuit I used on channel 3 (**fig. 6**) will cover TV channels 2 through 4 merely by substituting crystals and retuning the two trimmer capacitors. For operation on channels 5 and 6, it will be necessary to reduce L1 to 6 turns, L2 to 8 turns, and L3 to 2 turns. All other component values remain as in **fig. 6**. Similarly, the vestigial sideband filter shown in **fig. 7** may be tuned to cover TV channels 2 and 3. For operation on channels 4 through 6, L1 and L4 of **fig. 7** must be reduced to 3 turns, and L2 and L3 to 7 turns each. Table 1 will serve as a guide in selecting crystal frequencies. When the microtransmitter operating frequency is increased, output power will begin to degrade as the upper frequency limit of the integrated circuit is approached.

microtransmitter circuit

The basic circuit for generating 10 mW of stable double-sideband A5 with the LP-2000 microtransmitter

table 1. Video-carrier frequencies of lower vhf television channels.

channel number	video carrier frequency (MHz)
2	55.25
3	61.25
4	67.25
5	77.25
6	83.25

IC is shown in **fig. 6**. The circuit is divided functionally into three sections. J1 is the video input connector, which is driven by the standard composite video output signal from a TV camera or video tape recorder (typically 1 volt peak into a 72-ohm impedance). This video

*An experimenter-grade version of this microcircuit, the NA2000, is available for $9.95 from NASEM, Box AI, Cupertino, California 95014.

drive level is more than adequate to overmodulate the microtransmitter, hence the pad-and-trimpot combination at J1, which simultaneously matches the relatively high video input impedance of the IC to 72 ohms and allows the appropriate video level to be set.

Because of the number of stages employed, the lead to instability. I have achieved the greatest success by using a piece of PC stock only as a ground plane, positioning the components in space above it to minimize lead lengths. I call such a configuration a "breadspace," for breadboard suspended in space. (See the accompanying photographs.) This circuit would also

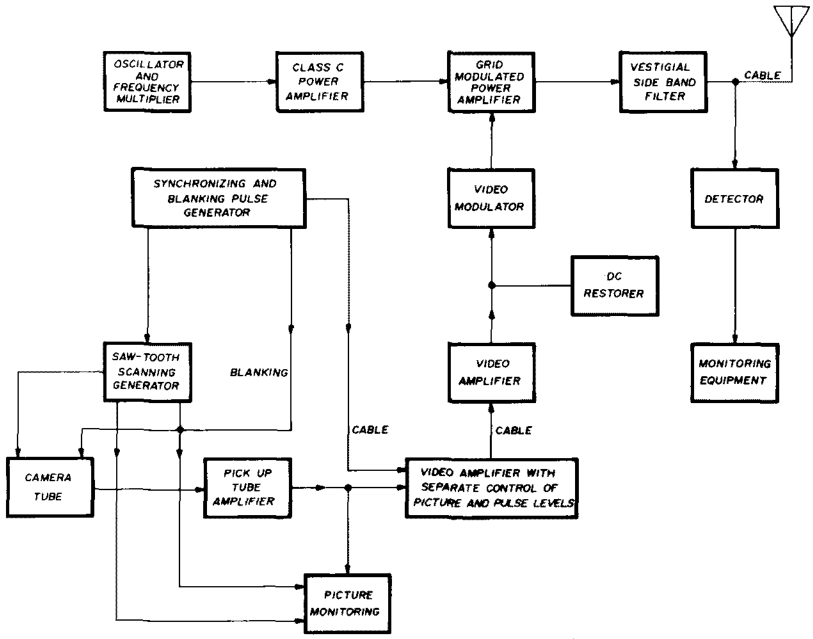


fig. 2. Simplified block diagram of a monochrome TV transmitter.

LP-2000 IC provides a considerable amount of rf gain in a rather confined space. Thus the circuit exhibits a strong tendency toward oscillations if precautions are not taken. It is advisable to let the physical arrangement of the schematic dictate the circuit board layout. As with all "hot" vhf circuits, short and direct wiring is a must. No printed-circuit artwork is provided, as the stray coupling between traces in a PC board would most likely lend itself well to isolated-pad construction, as described in recent articles.[3,4]

Parallel resonant circuits C1-L1 and C2-L2 tune the oscillator and amplifier stages respectively. Any coupling between them will obviously result in oscillations, or at least potential instability. Although the toroidal cores on which the inductors are wound tend to minimize stray coupling, the two inductors should nonetheless be

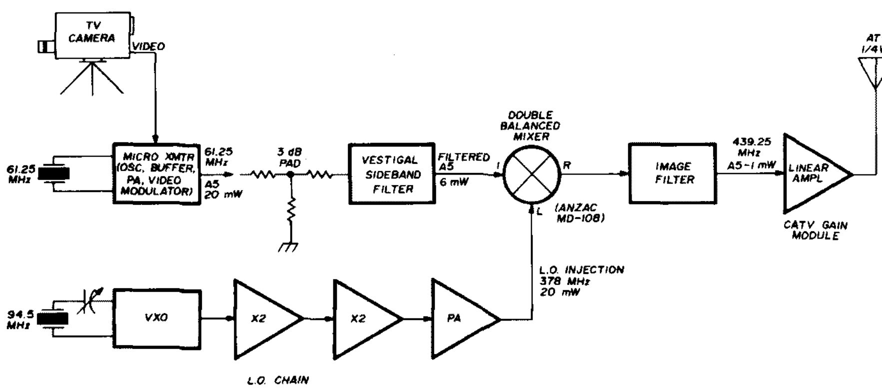


fig. 3. Block diagram of ATV transmitter for use in the 70-cm band.

oriented at right angles to one another as a precaution against oscillations. Although not attempted in the prototype unit, the use of shields positioned as shown by the dotted lines in **fig. 6** is a good idea. The 3-dB T pad between L3 and J2 not only keeps the power level within the requirements of the system but also provides a degree of isolation against instability that may occur from mismatching the output to its load.

microtransmitter tuning

A common amateur practice in tuning transmitting equipment is to adjust all tuned circuits for maximum indicated output power. As this circuit is potentially unstable, such an approach would be disastrous if applied to the microtransmitter. The resulting output signal could well contain a multitude of frequency components. If some of the output energy did indeed fall on the desired video carrier frequency, it would only be by coincidence. The best way to tune this circuit is with a spectrum analyzer. Trimmers C1 and C2 are tuned for maximum output on the desired video carrier frequency consistent with minimum spurious output. Tuning should be accomplished with video input connector J1 terminated into a 75-ohm resistor. Some interaction between the tuning of C1 and C2 will be noticed; repeated adjustments may be necessary.

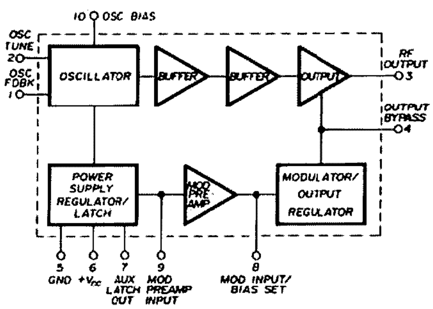
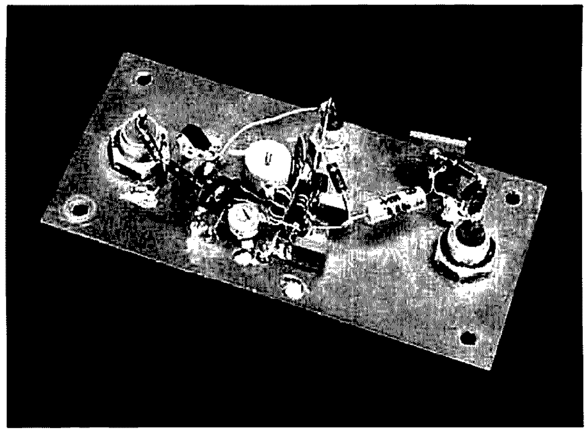


fig. 4. Block diagram of the LP-2000 transmitter IC.



Vestigial ATV system uses point-to-point wiring on PC chassis.

Since few amateurs have access to a spectrum analyzer, two alternative tuning methods are proposed. The first involves the use of a high-selectivity absorption wavemeter (or grid-dip oscillator in the absorption mode), *loosely* coupled to J2. Adjust C1 and C2 repeatedly for maximum indicated output on the desired video carrier frequency, then tune the wavemeter over its *total* frequency excursion to ensure absence of parasitic oscillations.

Those lacking an absorption wavemeter will probably have difficulty in adjusting this circuit. Nonetheless, a "last resort" tuning method may be attempted. Loosely couple J2 output into a TV receiver that is adjusted for reception at the channel for which the microtransmitter was built. Tune C1 and C2 until the resulting video carrier blanks the TV receiver screen. Now tune the receiver to all adjacent channels to detect any parasitic oscilla-

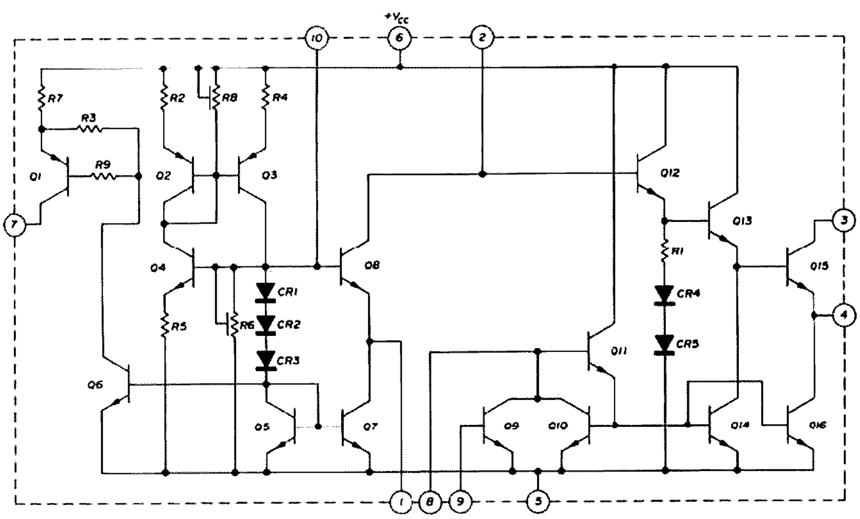


fig. 5. LP-2000 IC schematic.

tion. If any other channel is blanked, try again until output is noticed *only* on the desired channel.

The video level setting is best accomplished visually. After the rf adjustments are completed, loosely couple the rf output into a TV receiver. Connect a TV camera to J1 and scan a scene containing bright white level (a test pattern is ideal). Tearing of the horizontal synchronization will occur with the trimpot set for maximum video modulation. Back off on the video level until a stable sync is obtained, which will put the transmitter very close to the standard 12.5% ±2.5% modulation level for bright white. If the camera is properly adjusted, the 75% ±2.5% blanking level will fall into line automatically.

vestigial sideband filter circuit

The filter depicted in fig. 7, consisting of two critically coupled parallel resonant circuits with link coupling in and out, is the absolute minimum in circuit com-

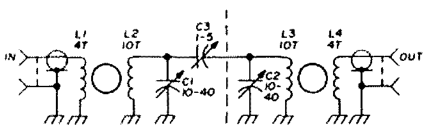
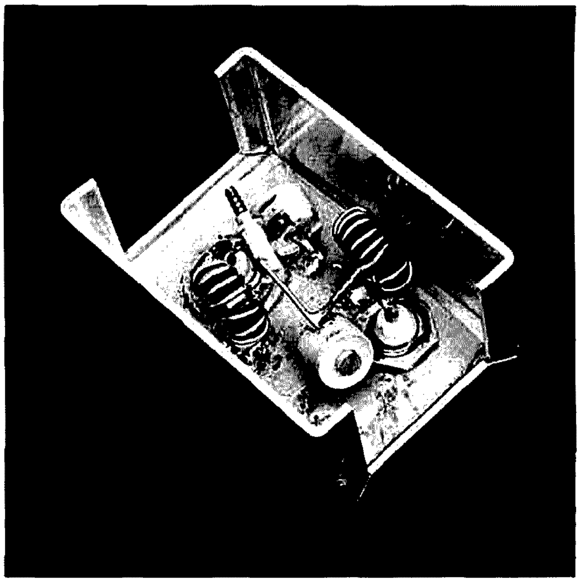


fig. 7. Vestigial sideband filter schematic for TV channels 2-3. All coils are wound with no. 18 (1mm) on Amidon T50-10 toroids. L1 is 4 turns; L2 is 10 turns; L3 is 4 turns; L4 is 4 turns. See text for coil data for channels 4-6.

plexity considered adequate for amateur vestigial sideband transmission. Attenuation of frequency components 2 MHz below the video carrier frequency, as seen in fig. 8, is 11 dB referenced to the passband midpoint. Similarly, the −3 MHz component is attenuated by 13.5 dB. If high-power ATV operation is anticipated, a greater degree of lower-sideband attenuation may be desirable, and two or more sets of resonator pairs may be cascaded. If multiple stages are used, stagger tuning may be necessary to maintain the required passband bandwidth.

As mentioned previously, the vestigial sideband filter may be modified for operation at different video carrier frequencies by modifying the number of turns on the toroids. As a general rule, skirt selectivity can be expected to degrade as operating frequency increases (due to a decrease in loaded Q). This suggests that cascaded



Construction of the vestigial sideband filter. All coils are wound on Amidon T50-10 toroid cores.

resonator pairs should be considered for operation at TV channels 5 and 6.

Construction of the vestigal sideband filter is far less critical than that of the microtransmitter module. The only precaution to be observed is adequate shielding of the filter assembly to prevent lower video sideband components from leaking around the filter and being radiated into following stages.

vestigial sideband filter tuning

As in the case of adjusting the microtransmitter module for optimum rejection of spurious output, properly tuning the vestigial sideband filter requires equipment not often available to the ATV experimenter. Thus

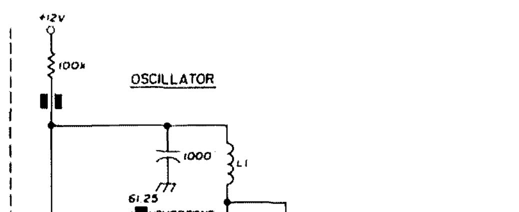
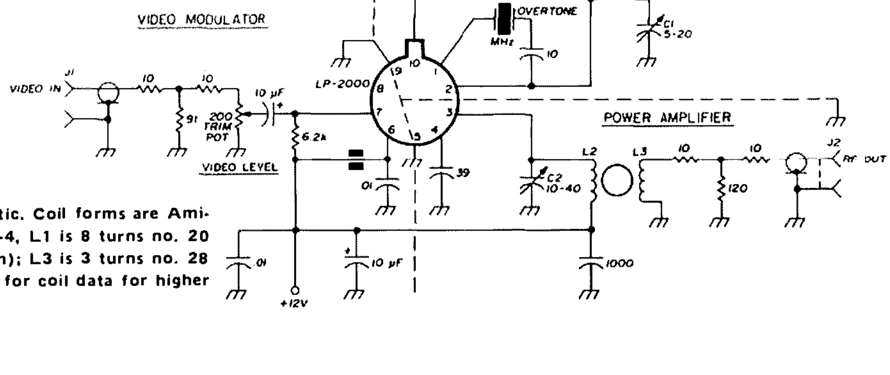



fig. 6. Microtransmitter module schematic. Coil forms are Amidon T25-10 toroids. For TV channels 2-4, L1 is 8 turns no. 20 (0.8mm); L2 is 10 turns no. 28 (0.3mm); L3 is 3 turns no. 28 (0.3mm) wound opposite L2. See text for coil data for higher channels.

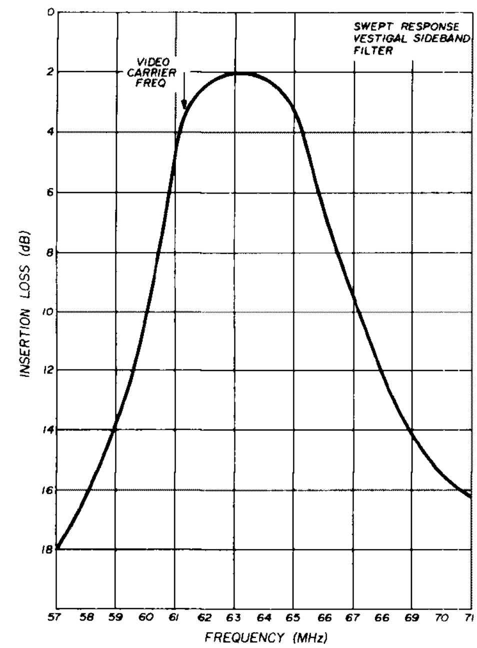


fig. 8. Vestigial sideband filter swept-frequency response.

in addition to the ideal approach, a compromise adjustment method will be outlined. Ideally, the filter should be adjusted on an rf sweep setup, as indicated in **fig. 9.** The procedure consists merely of adjusting C1, C2 and C3 of **fig. 7** repeatedly until the desired frequency response (that of **fig. 8**) is displayed on the CRT. The goals are a 5-MHz bandwidth, minimum passband ripple, and steepest possible lower-skirt selectivity with the video carrier frequency falling just at the knee of the lower-skirt rolloff. An application note from Hewlett-Packard[5] describes swept attenuation measurements in detail.

The filter passband can be adjusted manually using a stable rf signal generator, a vtvm with rf probe, and a 50-ohm coaxial feedthrough. Equipment is connected as in **fig. 10.** The signal generator is adjusted to 2 MHz above the desired video carrier frequency, coupling capacitor C3 adjusted to minimum capacitance, and C1 and C2 adjusted for a maximum indication on the vtvm. The filter will now be adjusted for minimum coupling (thus maximum Q) and will be resonant near the center of its passband. Next readjust the signal generator frequency to coincide with the video carrier frequency. The vtvm indication should drop off markedly because of the high selectivity and narrow bandwidth of the undercoupled resonators. The filter passband will widen if C3 capacitance is increased (because of tighter coupling), which will bring the video carrier within the lower skirt.

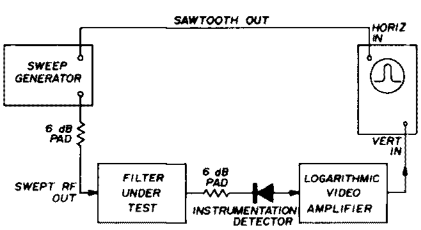


fig. 9. Setup for swept-frequency response measurement of filters.

The carrier-frequency attenuation, relative to midband power level, will be 1 to 2 dB when the voltage produced at the video carrier frequency (measured on the vtvm) equals 80 to 90% of the voltage indicated at mid passband. Acceptable vestigial sideband filtering will result under such conditions. Passband ripple and skirt selectivity can be examined readily by sweeping the signal generator manually in frequency and observing the vtvm.

conclusions

As rf spectrum space becomes increasingly scarce, vestigial sideband transmission will become the standard for ATV. A high degree of frequency flexibility can be maintained by generating a stable vhf television signal, filtering it to roll off the lower sideband, then heterodyning the resulting vestigial sideband signal to the transmission frequency. I hope the equipment described will be the first of numerous approaches to apply commercial standards to amateur television transmission.

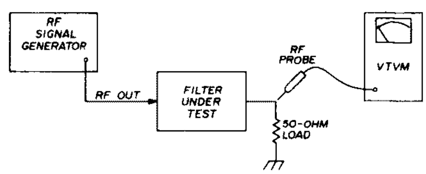


fig. 10. Setup for manual-frequency response measurement of filters.

acknowledgements

I wish to express my appreciation to Bob Hirschfeld, W6DNS, president of Lithic Systems, for his interest in developing amateur applications for his products. Thanks also go to Cliff Buttschardt, W6HDO, for encouraging me to try the LP-2000 even though, in his words, "it's a squirrelly chip." Once tamed, I found the device to be a fine choice indeed.

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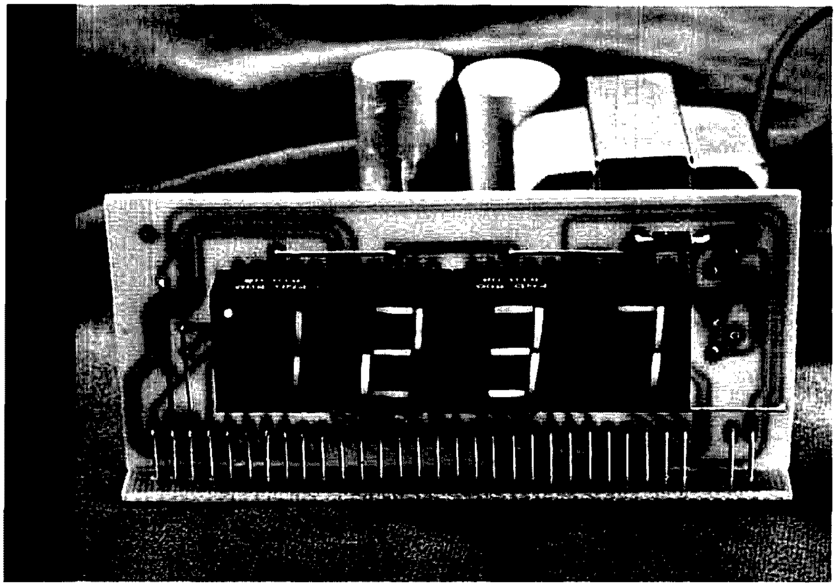


low-cost digital clock

New digital clock IC is designed for alarm clock-radio service and provides display drive, alarm and sleep-to-music in 12- or 24-hour formats

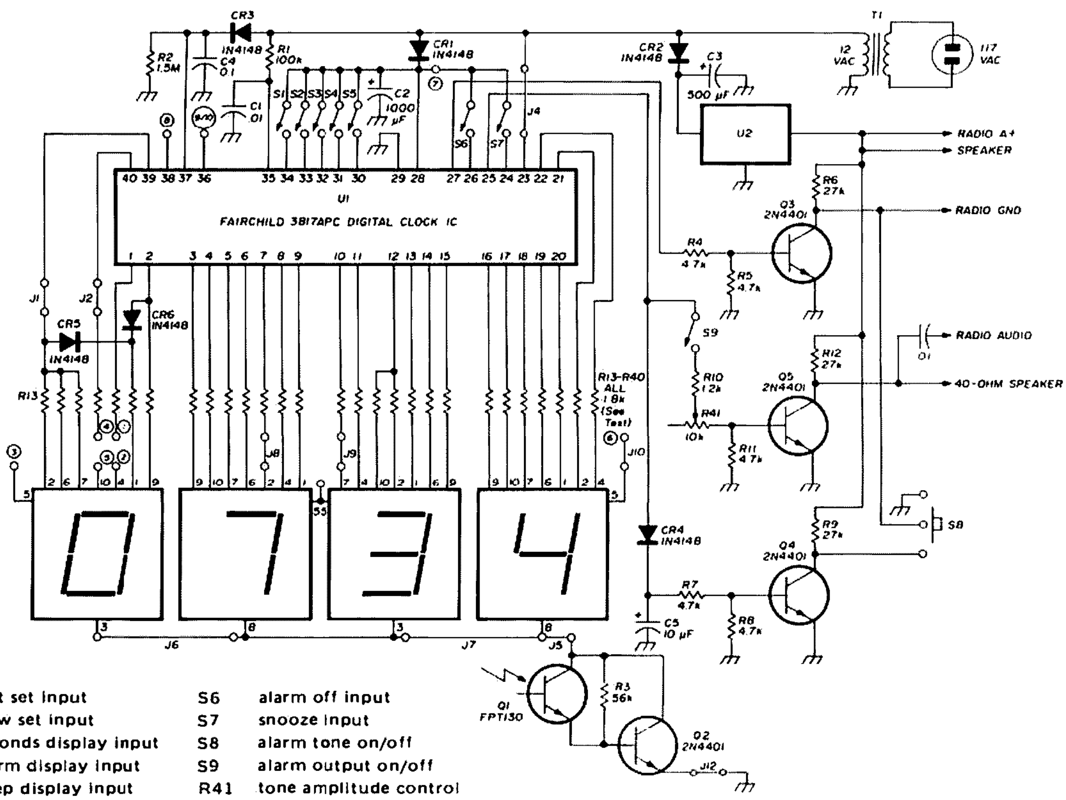
Fairchild Semiconductor has announced the 3817, an mos digital clock IC with full clock radio features. The direct drive offered by the 3817 IC allows the design of a simple, low-cost clock radio without the multiplex noise problem previously associated with mos clock circuits. The design described here capitalizes on this direct-drive capability and features the Fairchild FND500 LED display.

device description

The 3817 digital alarm clock is a monolithic mos IC which uses Isoplanar p-channel processing. The logic density thus achieved allows the incorporation of large output transistors for direct digit drive without making the overall chip size too large for low-cost, high-volume production. The 3817 is micro-programmable at the mask level to allow options such as alarm tone or dc at the alarm output pin without making major changes to the entire mask set. Four display modes are switch selectable (time, seconds, sleep and alarm) allowing the user to build several types of clocks and timers. Either a 50- or 60-Hz input may be used for the clock input, derived from either the power line with the simple RC filter shown or from an external timebase. Time display may be either 12-hour (with AM/PM indication) or 24-hour format. Outputs consist of display drive, alarm, and sleep to music (timed radio turn-off).

The FND500 is a 0.5 inch (13mm) high common-cathode LED display using a single diode per segment with a light pipe for diffusion. The digits may be horizontally stacked on 0.6 inch (15mm) centers for a compact display.

By Douglas R. Schmieskors, Jr., WA6DYW, 22065 McClellan Road, Cupertino, California 95014



S1	fast set input	S6	alarm off input
S2	slow set input	S7	snooze input
S3	seconds display input	S8	alarm tone on/off
S4	alarm display input	S9	alarm output on/off
S5	sleep display input	R41	tone amplitude control

fig. 1. Schematic diagram of the digital clock using the Fairchild 3817 digital clock IC. Circuit may be wired for 12- or 24-hour format, as discussed in the text. Pinouts of the 3817 are shown in fig. 2.

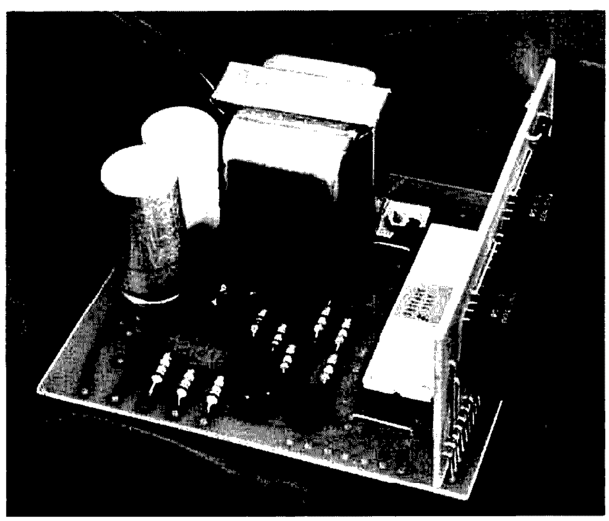
circuit description

Power supply. Three separate power supplies are actually used in the design shown in fig. 1. Diode CR1 and capacitor C2 provide V_{ss} and display power; CR2, C3 and U2 provide a regulated A+ for the radio; CR3 and C4 provide a "high" to the display blanking input of the 3817. Should a power failure occur, R2 discharges C4, the display blanking input goes low, and the display is blanked until power is reapplied. With the display blanked, the 3817 requires less than 4 mA to maintain the registers and this is provided by the charge on C2. U2 is a 7800-series IC voltage regulator with the output voltage and current handling capability determined by the requirements of the radio used. R1 and C1 form an RC filter to remove line transients which could cause false counting or device damage. The output of the filter is applied to the C_p input (pin 35) of the 3817, where an internal Schmitt trigger shapes the signal for further use.

Output drive circuits. Transistor Q3 and its associated resistors provide an active low output for timed radio turn-off after a user-selected interval of up to 59 minutes. This portion of the circuitry may be omitted in its entirety if the feature is not desired.

Diode CR4 and C5 rectify the alarm tone output for amplification by Q4, resulting in an active low output for timed radio turn-on when a coincidence is detected by the alarm comparators. Again, this portion may be omitted in its entirety if the feature is not desired.

Transistor Q5 and its associated components provide an alarm tone output at a level sufficient to drive a



Layout of the digital clock PC board. Three-terminal voltage regulator is not installed, nor is the phototransistor display control circuitry.

table 1. Display modes of the Fairchild 3817 digital clock IC.

selected display mode*	digit 1	digit 2	digit 3	digit 4
Time display	10s of hours and AM/PM	hours	10s of minutes	minutes
Seconds display	blanked	minutes	10s of seconds	seconds
Alarm display	10s of hours and AM/PM	hours	10s of minutes	minutes
Sleep display	blanked	blanked	10s of minutes	minutes

*If more than one display mode input is applied, the display priorities are in the order of Sleep (overrides all others), Alarm, Seconds, Time (no other mode selected).

40-ohm speaker with enough volume to wake even the soundest of sleepers. If a radio is used, this speaker should be omitted and 0.1-μF capacitor from Q5's collector to the radio's audio amplifier input should be installed. S9 is tone on/off and R41 controls the tone amplitude.

Control circuits. All control functions are implemented by applying V_{ss} to the appropriate pin (an internal pull down to V_{dd} through approximately 2 megohms is pro-

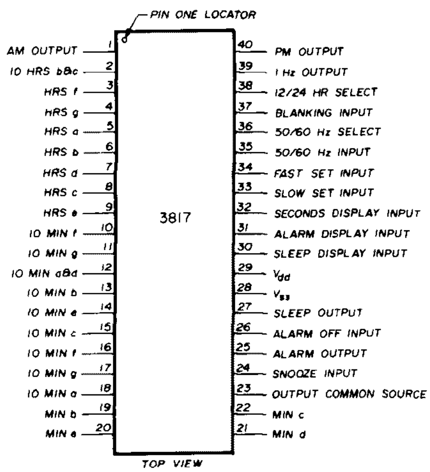


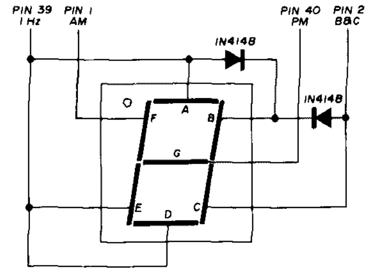
fig. 2. Pinouts of the 3817 digital clock IC.

vided.) Time of day is displayed in the absence of any of the following inputs:

Fast Set (pin 34) advances hours at a 1-Hz rate; *Slow Set* (pin 33) advances minutes at a 1-Hz rate; *Seconds Display* (pin 32) blanks the tens of hours digit and minutes and seconds are displayed on the remaining digits; simultaneous operation of *Seconds Display* and *Slow Set* displays seconds and holds the time counters (refer to tables 2 and 3 for a complete explanation of control and display functions).

Alarm Display (pin 31) temporarily defeats time-of-day display and causes the time for which the alarm is currently set to be displayed, along with the appropriate AM or PM indication when in the 12-hour format. *Alarm Set* is accomplished by simultaneous operation of *Alarm Display* and the appropriate setting input; the time-of-day setting is not disturbed by this operation.

fig. 3. Wiring of the tens of hours digit for the 24-hour format is accomplished on the PC board (fig. 5) by jumpers as discussed in the text.



Sleep Display (pin 30) blanks the hours digits and displays the minutes remaining until timed radio turn-off occurs. Operation of this input plus a setting input causes the sleep timer to decrement at the same rate at which time of day is set. When this input is activated, *sleep output* (pin 27) goes to V_{ss}; when the counter reaches 00 a latch is reset and the output goes low, Q3's collector goes high, and the radio turns off. The turnoff may also be accomplished at any time in the countdown by momentary operation of the *Snooze* input (pin 24).

Snooze inhibits the alarm output for 9 minutes, after which the alarm again sounds. The input may be used as often as desired during the 59 minutes for which the alarm latch is set.

Alarm Off (pin 26) resets the alarm latch, causing pin 25 to remain low and therefore silence the alarm. This momentary connection to V_{ss} also readies the latch for the next comparator output, causing the alarm to sound again 24 hours later. If no alarm output is desired for more than a day this input should remain at V_{ss}, so a spst toggle was used for this function. S9 is provided to silence the alarm tone while causing the radio to remain on for up to 59 minutes.

Digit drive circuits. Resistors R13 through R40 limit the output current of the 3817 to provide uniform display brightness and to prevent destruction of the output de-

table 2. Setting control functions for the 3817 digital clock IC.

selected display mode	control input	control function
Time*	slow	Minutes advance at 1-Hz rate
	fast	Hours advance at 1-Hz rate
	both	Hours advance at 1-Hz rate
Alarm	slow	Alarm minutes advance at 1-Hz rate
	fast	Alarm hours advance at 1-Hz rate
	both	Alarm resets to 12:00 AM (12-hour format)
	both	Alarm resets to 00:00 (24-hour format)
Seconds	slow	Hold (input to entire time counter is inhibited)
	fast	Seconds and 10s of seconds reset to zero without a carry to minutes
	both	Time resets to 12:00:00 AM (12-hour format)
	both	Time resets to 00:00:00 (24-hour format)
Sleep	slow	Subtracts count at 1 Hz
	fast	Subtracts count at 60 Hz
	both	Subtracts count at 60 Hz

*When setting time, sleep minutes will decrement at rate of time counter until the sleep counter reaches 00 minutes (sleep counter will not recycle).

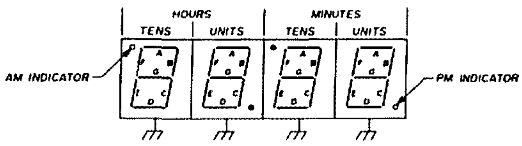


fig. 4. Arrangement of the FND500 common-cathode LED displays showing the AM and PM indicators. Note that the tens of hours and tens of minutes displays are inverted. In the 12-hour format the colon is provided by the unused decimal points included with the digits.

vices. The value of these resistors is determined by the formula

$$R = \frac{V_{ss} - V_f}{I_f}$$

Therefore, with a 12-volt rms transformer, R will be approximately 1800 ohms at 8 mA, as the forward voltage drop of GaAsP is about 1.6 volts and the 1000-μF filter capacitor will charge to the peak ac value.

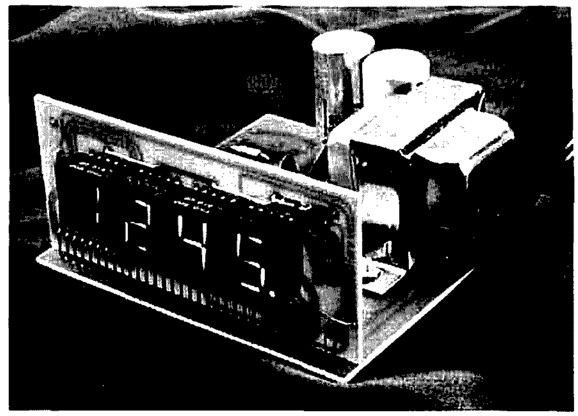
Wiring for the tens of hours digit in 24-hour time format is shown in **fig. 3** and is accomplished on the PC board (**fig. 5**) by jumper installation as follows:

12-hour operation: jumper points 1 and 3, 4 and 6, and omit J1

24-hour operation: jumper points 1 and 2, 4 and 5, and 7 and 8

In the 12-hour format, only, resistors R13 thru R40 may be omitted and replaced with jumpers if the following additional changes are made:

1. Replace J5 with a 5.1 volt, 1 watt zener diode with the anode oriented toward Q1 and Q2 collectors.



Front view of the low-cost digital clock shows installation of the separate readout circuit board.

2. Replace diode CR6 with a jumper. This maintains the display V common 5.1 volts above ground and moves a watt of power dissipation to the zener diode.

Display brightness control. Transistor Q1, a phototransistor, and Q2 control the voltage drop between the LED common cathodes and ground. R3 biases Q2 so that the display does not completely blank even in total darkness. 56 kilohms has been used with satisfactory results. Increasing the value will lower the minimum brightness with 100k being about the highest practical value. Q1 may be omitted and a 25k pot installed from V_{ss} to ground with the wiper connected to Q2's base, using the Q1 emitter pad for connection, if manual brightness control is desired. Q1, Q2 and R3 may be omitted and replaced with a jumper from Q2's collector to emitter for fixed maximum brightness.

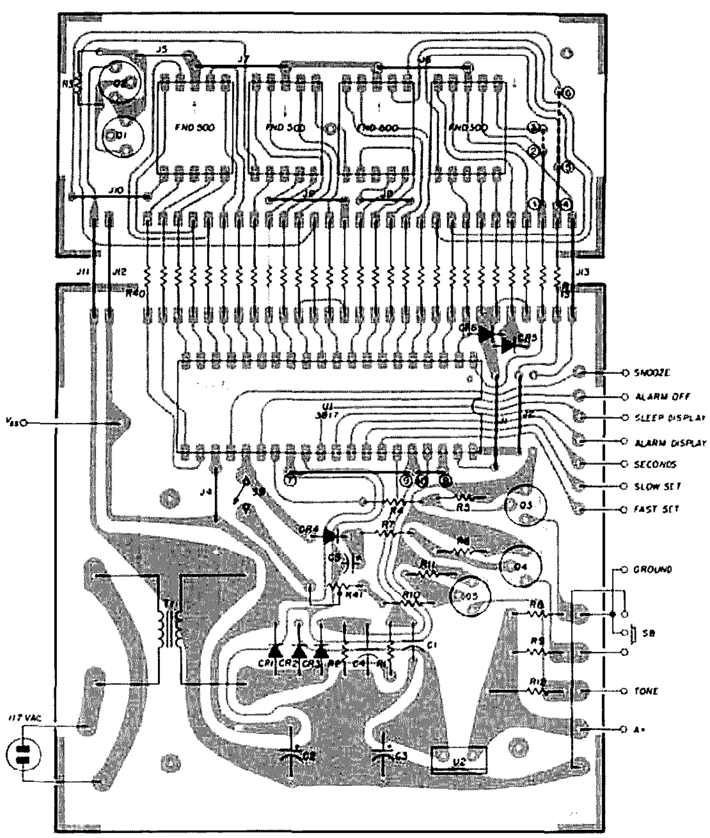


fig. 5. Component placement on the digital-clock PC board. A full-sized printed-circuit layout is shown in fig. 6.

Display. The tens of hours and tens of minutes digits (**fig. 4**) have been inverted in the display to provide an AM indicator and an acceptable colon from the otherwise unused decimal points included with the digits. This approach eliminates the use of discrete LED lamps for these functions. It should be noted that the manufacturer's designations of segments A thru G must be disregarded when a digit is inverted and the builder should re-define the segments as shown. The colon may be wired to the junction of CR2/C3 through a resistor in either the time display format, or, in 12-hour format only, it may be tied to the 1-Hz output thru a resistor one-half the value of that selected for R13 thru R40. This latter method will pulse the colon at a 1-Hz rate for

an activity indicator. An added benefit of this approach is that the colon brightness will then track display brightness since the 1-Hz output transistor is on the output common source bus rather than V_{ss}.

The AM or PM indicators are normally lit constantly (tens of hours digit in the 24-hour format); however, if V_{ss} drops below approximately 8 volts, the indicator will flash at the 1-Hz rate to indicate a potential display error. The indicator returns to a steady state after application of either setting input while in the time-of-day mode.

construction

Construction is very straightforward although normal handling precautions should be applied to the 3817 during construction. Small arrows on the display board foil side indicate the position of the orientation notches of the FND500 LED readouts.*

The single-sided PC board shown in fig. 3 is cut in two pieces at the dimensioning lines and R13 thru R40 with J11 thru J13 support the display board perpendicular to the main board as shown in the photograph. Operation at 50 Hz is selected by installing a jumper between points 7 and 10.

Table 3, a parts list, is included to provide a starting point for the builder. Few of the components shown are critical; in fact, resistance and capacitance values can be varied by 50% and more with no adverse effects, specified diodes can be replaced with virtually any diode with

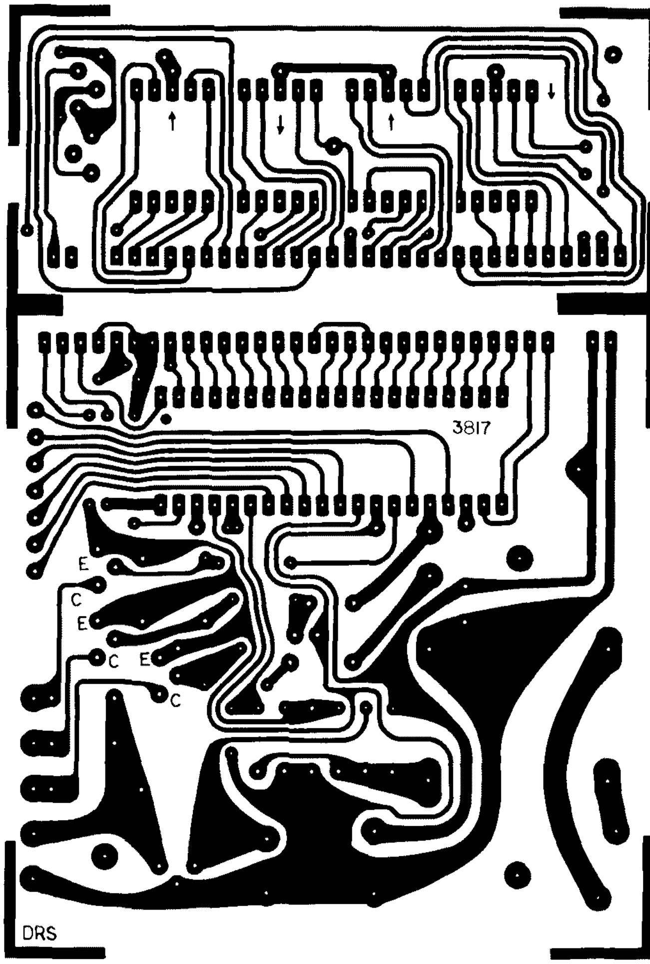


fig. 6. Full-size printed-circuit board for the digital clock.

a minimum of 10:1 front-to-back ratio, and the transistors may be virtually any available npn type.

table 3. Parts list for the digital clock.

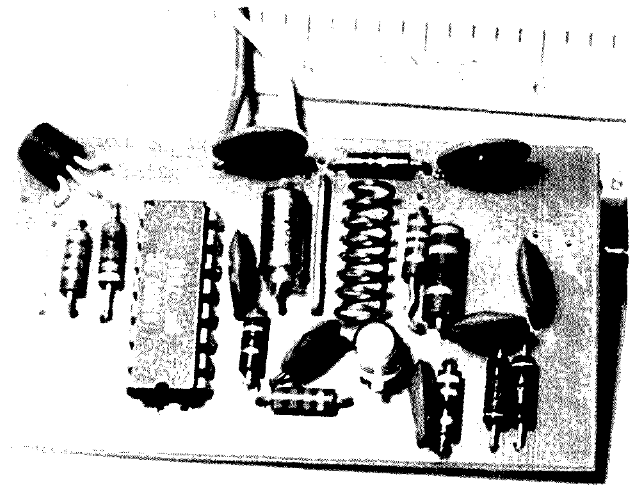
qty	part	description
1	C1	0.01 μF, 25 WVdc disc ceramic
1	C2	1000 μF, 25 WVdc electrolytic
1	C3	500 μF, 25 WVdc electrolytic
1	C4	0.1 μF, 25 WVdc disc ceramic
1	C5	10 μF, 25 WVdc electrolytic
6	CR1-CR6	1N4148
1	Q1	FPT130 phototransistor
4	Q2-Q5	2N4401 npn transistor
1	R1	100k, 10%, ¼-watt
1	R2	1.5 megohm, 10%, ¼-watt
1	R3	56k, 10%, ¼-watt
5	R4,R5,R7, R8,R11	4.7k, 10%, ¼-watt
3	R6,R9,R12	27k, 10%, ¼-watt
1	R10	1.2k, 10%, ¼-watt
28	R13-R40	1.8k, 10%, ¼-watt (see text)
1	R41	10k potentiometer
6	S1-S5,S7	spst pushbutton switch
2	S6,S9	spst toggle or slide switch
1	S8	spdt toggle or slide switch
1	T1	12 Vac secondary transformer, rating as required by radio
1	U1	Fairchild 3817APC digital clock IC
1	U2	78Lxx or 78Mxx voltage regulator (voltage and current determined by radio requirements)

*Printed-circuit boards and semiconductors for the digital clock are available from Circuit Specialists Company, Post Office Box 3047, Scottsdale, Arizona 85257: set of two circuit boards, \$4.50; Fairchild 3817APC clock IC, \$6.50; FND500 LED readouts, \$3.50 each; MPSA70 transistor (2N4401 replacement), 32¢ each.

conclusion

An attempt has been made to illustrate a minimum-cost but full-featured clock radio design which can be scaled down to a simple desk clock if so desired by the builder. The usual multiplexing noise associated with electronic digital clocks is eliminated by the direct drive approach, while overall circuit cost and complexity is reduced. The 3817 IC should find a home in many other applications such as automobile clocks (using a crystal and 12-state cmos divider for time-base generation and the blanking input to kill the display in ignition off conditions). Photography timers, appliance timers, industrial controllers, and digital stopwatches are other potential uses. Other common-cathode displays such as the FND70 may be used in place of the FND500 shown, or liquid crystal, neon, or fluorescent display may be substituted at some cost sacrifices. The 3817, FND500, and related data sheets may be obtained from franchised Fairchild distributors.

ham radio



vhf prescaler

for digital frequency counters

A sensitive circuit that will extend your counter's range to 300 MHz

As easy-to-build, high-performance vhf prescaler can be added to a high-frequency digital counter to extend its useful frequency range to 300 MHz. Such a prescaler is described here with my experiences in adding it to a homebrew counter. The prescaler is based on the Fairchild 95H90 and a series of articles by W6PBC.[1,2,3] These articles contain excellent material for those interested in obtaining the ultimate performance from the 95H90 prescaler. The unit described here can be built for around $25.

circuit

The circuit (fig. 1) is based on W6PBC's work and information in Fairchild 95H90 data sheets.[4] A low-impedance input amplifier uses a 2N5179 transistor to provide good gain at vhf. Two back-to-back 1N914 diodes protect the input from overload. The 22-ohm resistor and two 0.05-μF capacitors isolate the input amplifier to prevent oscillation. The amplifier is coupled to the 95H90, which also has a low input impedance. The 95H90 operates best with low input impedances hence the 68- and 200-ohm input bias resistors.

Decoupling the 95H90 from the power supply is accomplished with the 0.01 and the 2-20 μF capacitors. A 2N5771 couples the 95H90 output to TTL counter inputs. Most counters with amplified inputs can be

By Marion D. Kitchens, Jr., K4GOK, P.O. Box 183, Haymarket, Virginia 22069

operated by connecting pin 8 of the 95H90 to the counter input through a 0.01-μF capacitor. Both arrangements are shown in the schematic and the parts placement drawing, fig. 2. Input sensitivity was not measured but should be around 15 mV at 100 MHz and about 100 mV at 260 MHz, according to W6PBC's data.

construction

Construction is simple. Just mount all parts, except R_x, onto the PC board and solder. (R_x is discussed later.) A few points about construction should be made, however. The 2-20 μF decoupling capacitor should be the smallest physical size you can obtain. The PC board is laid out for ¼-watt resistors, although if you really work at it you can install ½-watters. The ¼-watt resistors are preferred. Note that pin 14 of the 95H90 is floating; no connection should be made to it.

The RG-174 coax is held in place (strain relieved) by placing short loops of wire over the coax and soldering them to the PC board. Fig. 2 shows component placement. A photo of the circuit board is shown without the heatsink in place and with temporary wiring. The

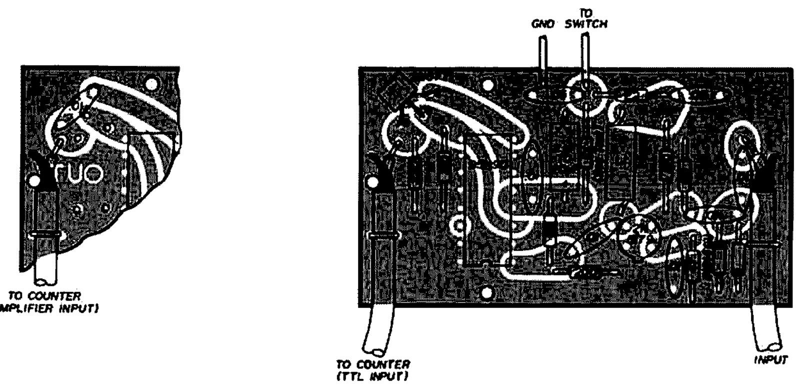
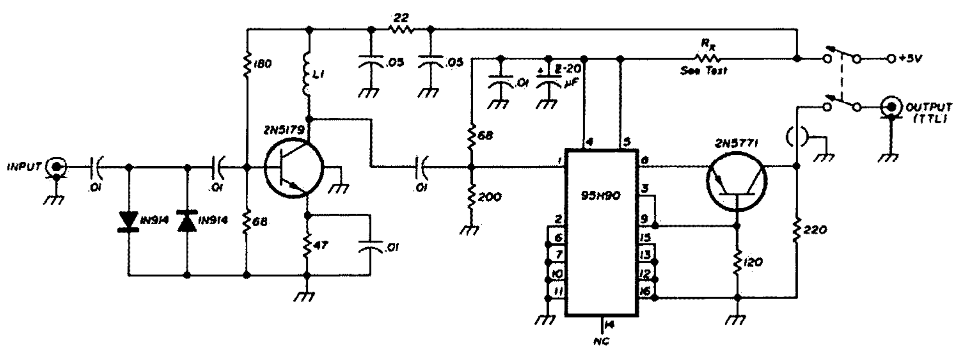


fig. 2. Full-size board showing component placement.



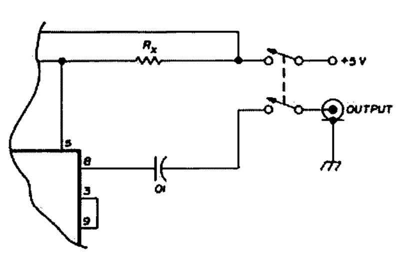


fig. 1. Vhf prescaler schematic using the Fairchild 95H90 (after W6PBC). L1 is 8 turns no. 26 (0.4mm), 5/32" (4mm) diameter, 5/8" (16mm) long.

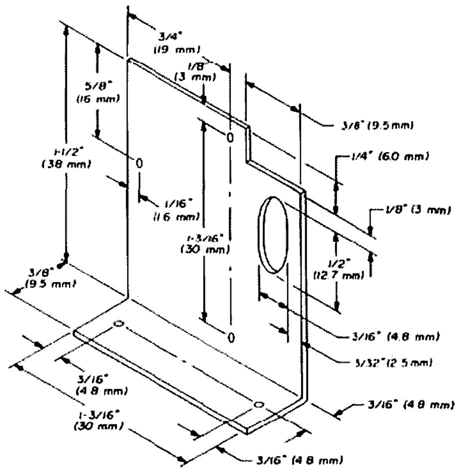
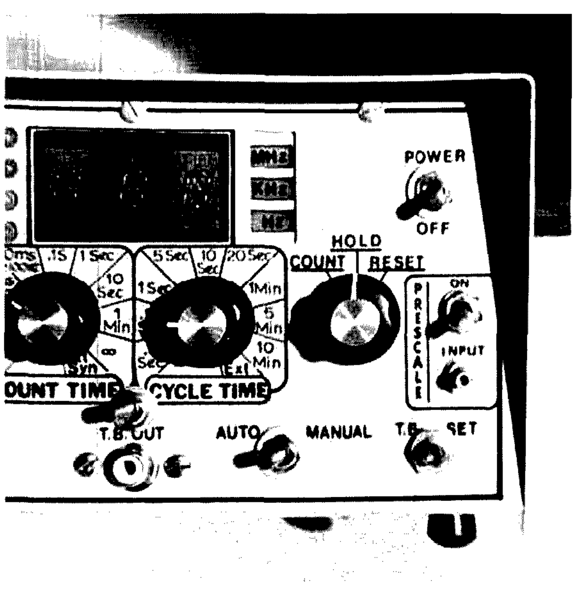


fig. 3. Heatsink is made from 0.03-inch thick (1.0mm) aluminum.

2N5771 should be installed with its flat side down against the PC board so it will clear the aluminum heatsink.

A heatsink (fig. 3) is strongly recommended although not absolutely necessary. Fairchild data sheets indicate the 95H90 maximum count frequency depends on the IC's temperature. About 750 kHz is lost for each °C rise in temperature, for near-room temperatures. My prescaler frequency limit was around 270 MHz without the heatsink, but went to 300 MHz with it. The signal source, a grid-dip oscillator, was limited to 300 MHz so I don't know if my prescaler will go higher or not. The heatsink was used to mount the prescaler in the cabinet. **Fig.** 4 shows a full-size etched board layout for the prescaler.

Only external evidence of the vhf prescaler in K4GOK's homebrew frequency counter is the separate prescaler input connector and on-off switch on the right side of the front panel.



supply voltage

Individual 95H90s have a "best" supply voltage that results in maximum count frequency. The best voltage for most 95H90s is 4.85 V according to W6PBC's data. My homebrew counter power supply provides 4.85 volts (how lucky can you get?), so when the prescaler was wired directly to the power supply for testing a 300-MHz count frequency was obtained. However, when the permanent installation was made, the maximum count frequency was only 150 MHz. After many hours of searching I found the 95H90 voltage was only 4.60 V. I was surprised to find a 2-amp in-line fuse produced a 0.1-volt drop. The other 0.15-volt drop was across a switch located between the power supply and the counter. This total 0.25 volt drop caused no problems with the basic counter but sure played havoc with the prescaler.

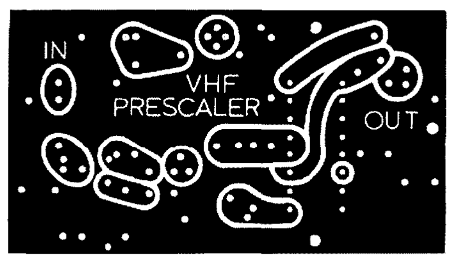


fig. 4. Full-size etched board layout.

The 95H90 draws 100 to 150 mA so a value for R_x in the 1-ohm range will provide 4.85 volts from a 5.0-volt source. Individual 95H90s will draw different currents, so R_x is best determined by trial and error. Tack in a trial resistor and measure the maximum count frequency until you're satisfied.

conclusion

The prescaler was easy to build and operate. It should be useful for vhf enthusiasts since it covers 50, 144 and 220 MHz with good sensitivity. No tricky, fussy or unstable circuits are involved. The vhf prescaler is a very worthwhile addition to all digital counters.

references

1. F. E. Emerson, W6PBC, "Advanced Divide-by-Ten Frequency Scaler," *ham radio*, September, 1972, page 41.
2. F. E. Emerson, W6PBC, "Circuit Improvements for the Advanced Frequency Scaler," *ham radio*, October, 1973, page 30.
3. F. E. Emerson, W6PBC, "Comments on Frequency Scaler," *ham radio*, November, 1973, page 64.
4. "95H90 Very High Speed Divide by 10/11 Prescaler," Fairchild Characteristics And Applications Data Sheet, Fairchild Semiconductor, July, 1973.

ham radio

50 years of television

The first public demonstration of television was given fifty years ago although experimenters have been interested in transmitting visual images for more than 100 years

Last year marked approximately the 50th anniversary of television, a media which began amidst an array of flickering neon lamps and a whirling disc in 1925, when C. Francis Jenkins transmitted a live silhouette of a moving windmill from his workshop in Anacostia, Maryland, to the Navy Department in nearby Washington, DC. Later that year Jenkins gave his first public demonstration of the radio transmission of live images (which he called "radiovision") and film ("radiomovies").

In a conference at the Department of Commerce on May 29th, 1925, the authorities decided to allow amateurs to transmit pictures and facsimiles on any wavelengths for which they were licensed,[1] but there is no record of any amateur television transmissions until many years later.

the early years

The transmission of visual images goes back one-hundred years, to 1875, when George Carey in Boston used the system of fig. 1 to simultaneously transmit each separate picture element by wire. This followed the discovery in 1873 by Lewis May, a British telegrapher, of the photoconductive properties of selenium. The principle of rapidly scanning each picture element in succession, line by line, was proposed in 1880 by Maurice Leblanc of France and led to one of the first television patents which was issued to Paul Nipkow of Germany in 1884. The distinctive feature of the Nipkow system was the use of a spinning disc, with a spiral array of holes near its outer edge, to disassemble the image into a series of dots, and a similar disc at the receiving end to reassemble the picture (fig. 2). Until the advent of all-electronic image scanning in the 1930s, all workable television systems depended on some form or variation (mirrored drums, lensed discs, etc.) of the sequential scanning system examplified by the Nipkow disc.

The sequential reproduction of visual images is feasible only because the human visual sense displays a persistance of vision -- the brain retains the impression of illumination for about 100 milliseconds after the source of light is removed. If the image-making process occurs within less than 100 milliseconds, the eye is unaware that the picture has been assembled piecemeal, and it appears that the whole viewing screen is continuously illuminated.

Although selenium was used by all the early television experimenters, it had one serious handicap: slow response to changes in light. The discovery of a potassium hydride coated cell in Germany in 1913 improved sensitivity and the ability to follow rapid changes of light,

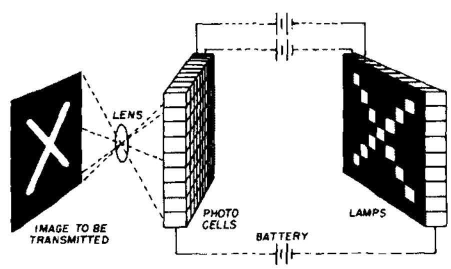


fig. 1. Image transmission system of 1875. At the transmitting end light is converted into electrical energy which is used to energize a lamp at the receiving end. Since the output of each selenium cell must be individually connected to a corresponding lamp at the receiving end, a large number of wires is required.

By Jim Fisk, W1DTY, and Dave Ingram, K4TWJ

but experimenters were still limited by the slow response of incandescent lamps. This was solved by the invention of the neon gas-discharge tube by D. M. Moore in 1917.

The application of the cathode-ray tube for television reception was first proposed by Boris Rosing in Petrograd in 1907, but development of his patent was retarded by the lack of suitable photo cells and electronic amplification. However, he did succeed in transmitting and reproducing some crude geometrical patterns. In 1908, Alan Campbell-Swinton, a Scotsman, outlined a method that is the basis of modern television when he proposed the use of a magnetically-deflected CRT at both the camera and receiver. Although his idea couldn't be translated into workable hardware because he lacked suitable amplifiers, his proposed mosaic-screen image pickup tube was remarkably like the iconoscope invented by Vladimir Zworykin some fifteen years later.

Because of the lack of suitable amplifiers, television experimenters continued to work with mechanical television systems and, in 1922, using his own version of Nipkow's disc, C. Francis Jenkins transmitted a still picture from one room to another. The next year he received nation-wide attention when he sent a recognizable picture of President Harding by wireless from Washington to Philadelphia.

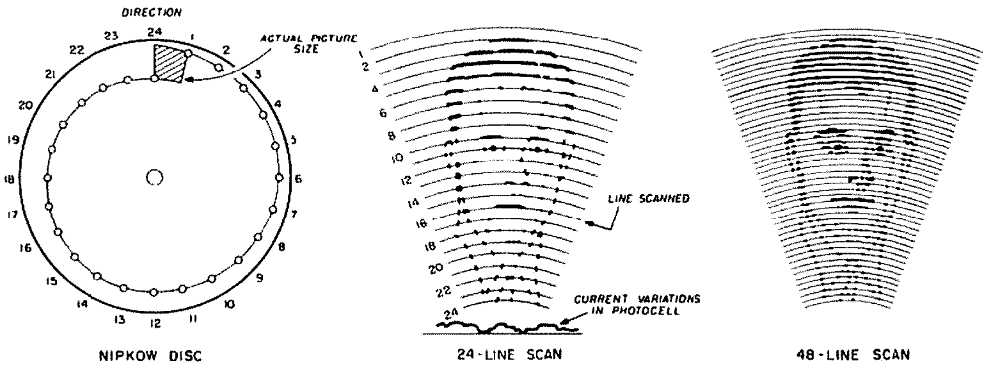
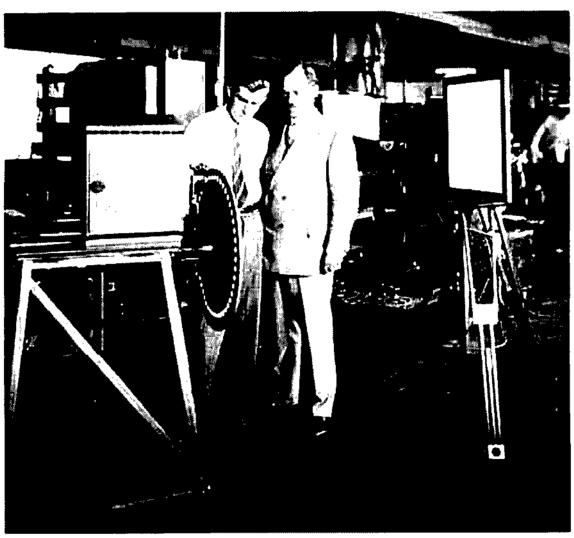


fig. 2. Spiral hole layout in a 24-line Nipkow scanning disc is shown at left. The shaded area represents the size of the reproduced image. Enlarged view of an image produced by the 24-hole scanning disc in center shows poor resolution of 24-line scanning. At right is the same image produced by 48-line scanning disc. Resolution is improved but is still crude compared to modern standards.

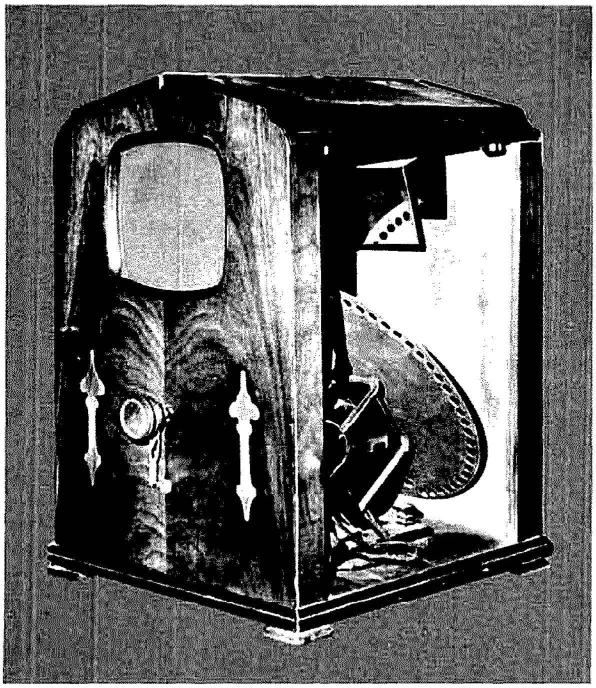


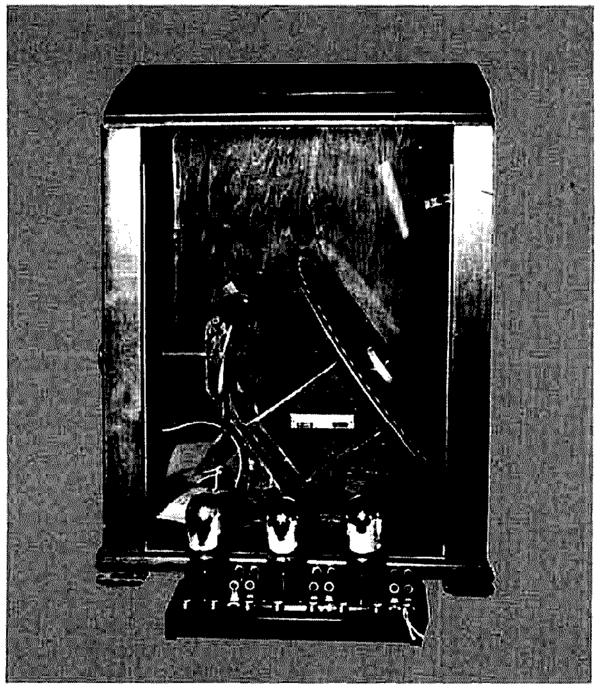
Television pioneers Dr. E.F.W. Alexanderson (right) and Ray D. Kell examine a lensed 48-hole Nipkow disc used in TV experiments in 1927. This disc is now in the collection of the Antique Wireless Association Museum in East Bloomfield, New York. (Photo courtesy RCA)

In the Jenkins system a disc with 24 (later 48) apertures was rotated at 2000 rpm by a motor whose speed was varied until it was synchronized with a similar setup at the transmitting end.[2] A neon tube was positioned behind the receiving disc and connected in place of the receiver's earphones which, in the broadcast sets of the 1920s, were connected between the audio output tube's plate and B+ supply. A piece of ground glass or thin wax paper was placed in front of the neon tube to diffuse the light. Motor speed was difficult to regulate, and since exact synchronism was required for good image reproduction, copying a picture off the air was something of a challenge. The pictures were usually about two inches (51mm) square although many viewers (Jenkins called them "Lookers in") used a magnifying glass to enlarge this area to 5 or 6 inches (13 to 15cm) square.

All of the mechanical systems, however, suffered from poor definition and flickering. Swinton and others had pointed out that at least 100,000 and preferably 200,000 elements were required lor good quality and definition on a screen of reasonable size. John Baird of England gave the first true demonstration of television in 1926 by transmitting moving pictures in *halftones* using 30 lines, scanned 10 times per second. However, since the number of elements is approximately equal to the square of the number of lines, Baird's 30-line system was far from adequate — 300 lines being more nearly the minimum.

Far-sighted planners at American Telephone and Telegraph, Westinghouse, RCA and General Electric saw the commercial possibilities of television in the mid-20s, and in April, 1927, AT&T set up the first long-distance telecast in the United States when then Secretary of Commerce Herbert Hoover spoke from a makeshift studio in Washington, DC, and sight and sound were received over the telephone circuits in New York City, 200 miles away. In the considerable publicity given to the AT&T transmissions the term "television" was used and soon came into widespread use as applying to any form of visual broadcasting. "Television," of course, means transmission over wire, and although you cannot argue with established usage, Jenkins stubbornly continued to call the new medium "radiovision" in his magazine articles and advertising.



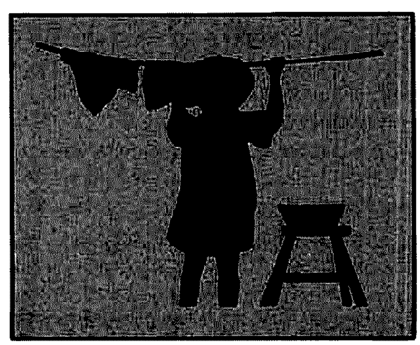


Jenkins Radiovisor from the collection in the AWA Museum. In this set the image was transmitted through a rotating Nipkow disc to a mirror and reflected onto the ground glass viewing screen. (Photographs by W2BWK)

In the summer of 1928 the Federal Radio Commission issued experimental television licenses to Jenkins Laboratories in Washington (W3XK) and to the General Electric Company in Schenectady, New York (W2XCW). Jenkins began broadcasting radiomovies on a regular schedule on July 2nd, and a month later he reported that "one hundred or more had finished their receivers and were dependably getting our broadcast pictures . . . " Hugo Gernsback's magazine, *Television*, regularly reported new developments and published construction articles for amateurs eager for information and was packed with advertising for television kits, parts, discs and neon lamps. Gernsback also owned a pioneering radio station in New York, WRNY, which broadcast live pictures for the first five minutes of every hour along with simultaneous sound which was broadcast by another local station.

In September, 1928, General Electric telecast the first *live* video drama from W2XCW in Schenectady, an old play called "The Queen's Messenger," selected primarily because it had only two characters. The accompanying sound was transmitted by WGY, GE's a-m broadcast station. This created a lot of excitement in the press, but Dr. Ernst Alexanderson, who directed much of the television development at GE, cautioned that the program was experimental and didn't mean that television was yet ready for public consumption.

Indeed it wasn't. Although *QST* devoted more space to television in 1928 than it did to radiotelephony, amateur interest in the crude radiovision systems of the day waned quickly, and the topic received little coverage in *QST* in 1929. It wasn't until eight years later and the



Silhouette as broadcast by the Jenkins Laboratories in 1928. The image size shown here is the approximate size of the image seen by "lookers in."

development of cathode-ray television systems that *QST* expressed renewed interest in the subject.[3,4]

Aside from the very crude pictures of the disc-television systems which made it difficult, in the words of one wit, " . . . to tell the difference between an opera singer and her poodle," the big problem was synchronization. The utility companies tried to maintain 60 Hz, but there were no unified power grids as we know them, so synchronization was always slow and laborious, and often impossible. To quote ARRL's Percy Maxim, W1AW, " . . . for about half a second, I actually had a picture. It flickered and it was fuzzy and foggy, and about the time I was wondering why they picked on a cow to televise, it suddenly dawned on me that it was a man's face I was looking at. Then I lost synchronism and my man disappeared into a maze of badly intoxicated lines . . ."[5]

By 1929 a total of twenty-six television stations were licensed by the Radio Commission, although few of them broadcast with any regularity. Jenkins, however, increased the power of W3XK and started work on a plant in New Jersey to build "radiovisors." In 1930 he petitioned the Radio Commission to commercialize television using his 60-hole disc system, but the request died without action when a Commission engineer said the mechanical system was "an absorbing field for the experimenter but not ready for entertainment." The major corporations – including GE, RCA and Westinghouse – echoed the Commission's view.

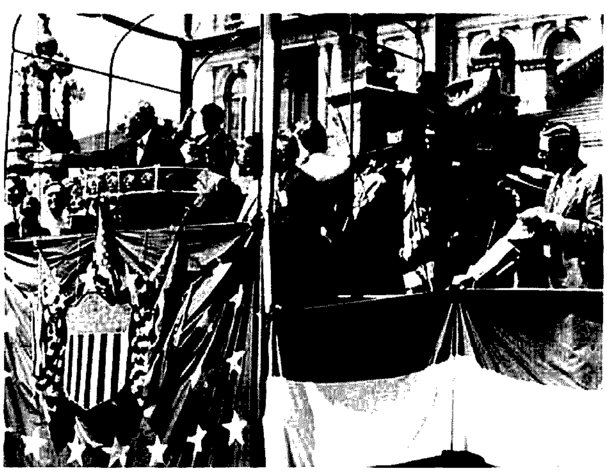
The images, as seen in the receiver, were small and extremely crude. In addition, the pickup camera was fixed so the subject had to be brought to it, and the transmission of a person's head and shoulders strained all the resources of the scanner and transmitter. Obviously, a telecast within such technical limitations could have little entertainment value.

Nevertheless, continuing research resulted in increasing the number of scanning lines to about 180 lines per picture, and later, to 240-line images, all generated by mechanical methods. The increased image details forced higher and higher speeds in the mechanical parts, until engineers despaired of ever presenting an image of fine detail by mechanical scanning methods.

By this time news of all-electronic image-scanning systems were beginning to reach the hobby magazines, and the days of the whirling discs were numbered. In 1932 Jenkins terminated his broadcasts and was taken over by the Deforest Radio Company, which itself later drifted into bankruptcy.

The first home television reception took place in 1927 at Dr. Alexanderson's home in Schenectady. A television system developed by Alexanderson and his co-workers was used for public broadcasts in 1928, the year this photo was taken. The television screen is in the small square at eye level. (Photo courtesy GE)

General Electric televised the first remote television broadcast in Albany on August 22, 1928, when Governor Alfred E. Smith accepted the Democratic nomination for President. (Photo courtesy GE)



1925 re-activated

As a nostalgic special interest project, K4TWJ is planning a re-activation of 1925 style television. This will be a project designed so that anyone can join in the fun at minimum expense (less than ten dollars, depending on your junk box). Alan Smith, W8CHK, and Dave Ingram, K4TWJ,* will work together to supply information to interested parties. Alan will have the disc patterns, detailed sketches, and instructions available via mail; Dave will distribute cassette tapes of TV signals (include return postage).

The tapes will be handled on an exchange basis. Originally, a one-time transmission of 1925 TV signals was planned for 20 meters. These TV signals sound a bit like someone tuning up (a 1000 Hz note) and occupy little bandwidth.* Briefly, K4TWJ's request to the FCC was for a one time, three-minute TV transmission which

*Alan Smith, W8CHK, 3213 Barth Street, Flint, Michigan 48504.

Dave Ingram, K4TWJ, Eastwood Village, No. 604N, Rt. 11, Box 499, Birmingham, Alabama 35210.

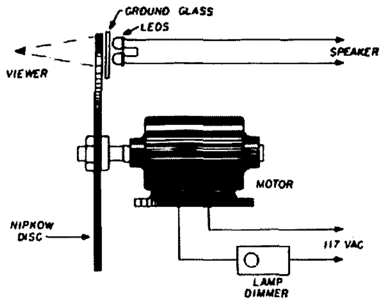


fig. 3. System used by K4TWJ to re-activate 1925-style television.

would be included in a roundtable discussion among interested parties, and would be abandoned if QRM was evident. The FCC turned thumbs down on the request, contending that the mode was outdated, the transmission would create unnecessary interference, and suggested a nationwide telephone hookup. Mailing tapes was the alternative.

Fig. 3 shows K4TWJ's modern day equivalent setup for reproducing 1925 style TV. An ac motor such as an old fan or phono motor is used to rotate the scanning disc. A commercial light dimmer is placed in series with the motor for sync/speed control of disc. The speed will later be adjusted to approximately 1700 rpm. Light-emitting diodes are used to replace the neon glow tube used in the early systems. It is suggested that three or four LEDs be used and positioned to form a square picture area. Wire the LEDs in parallel and connect them to your receiver or tape recorder speaker. A piece of ground glass (grind in a mixture of turpentine and sand) is placed between the LEDs and the disc to produce a picture area. This area will be approximately 1/2 inch (13mm) square, depending on the LED's size.

Harry Mills, K4HU, and Alan Smith, W8CHK, are to be thanked for their assistance with this project. Harry's experiences in actually receiving the original TV transmissions from GE's station in Schenectady was the final push needed to get it going. His assistance in locating information on disc TV systems in old engineering texts was very helpful. Alan, W8CHK, heard of the project and offered to help with copying and mailing. Their assistance is greatly appreciated.

electronic image scanning

In 1923 Dr. Vladimir K. Zworykin, a former student of Boris Rosing in Petrograd, was granted a patent on a system for the "cell storage of light" that was to become the basis of modern television. A year later he demonstrated a crude tube, which he called the *iconoscope*,† that scanned a scene electronically.

In the iconoscope (**fig.** 4) the *external image* is focused on a mica plate which is covered on one surface by millions of photosensitive particles, each insulated from the other (called a mosaic plate). The other side of the plate has a thin, deposited metal coating (called the signal plate) so each photosensitive particle forms one plate of a miniature capacitor. When a scene is focused on the mosaic plate, each of the particles develops a positive charge which is proportional to the amount of light falling upon it. When the photosensitive mosaic is scanned by an electron beam, the beam discharges each of the particles, in turn, and creates a small electric current which is picked off the signal plate and amplified.

Although the iconoscope was used in nearly all the early electronic television systems, secondary electron emission generated undesired outputs which had the effect of producing uneven shading. As a result, the reproduced image had large areas with varying brightness levels which were not contained in the original scene. This spurious shading signal is often called *dark-spot* shading because it can be generated when the mosaic plate is not illuminated. The spurious shading signal is inherent in the iconoscope camera tube but is minimized by using low values of beam current (at the expense of camera efficiency).

At about the same time Zworykin was working on his iconoscope, Philo T. Farnsworth was working independently toward an electronic scanning system somewhat along the same lines. However, while the iconoscope is based on electron storage, Farnsworth's *image dissector* camera tube may be considered an instantaneous scanner. The image dissector, shown in fig. 5, consists of a flat photosensitive cathode located at one end of the tube. The light from the scene is focused on the cathode, and electrons are emitted in proportion to the amount of light striking it at any one point.

The electrons emitted from the cathode are forced to move down the tube by high positive voltages applied to

K4TWJ and the Nipkow disc he built to reactivate 1925-style TV.

*There is a close resemblance between 1925 TV signals and modern slow-scan television. Both signals use audio tones which require minimum rf bandwidth. Slow-scan TV, however, is infinitely more stable and has much higher definition. If 1925 TV sounds interesting, you are invited to investigate sstv. Any active slow-scan operator will be glad to get you started. Also, the weekly SSTV Net which meets on 14.230 MHz on Saturdays at 1800 GMT welcomes inquiries.

†From Greek *icon*, "image," and *scope*, "to observe."

Dr. Vladimir K. Zworykin, inventor of the iconoscope, all-electronic "eye" of the television camera. (Photo courtesy RCA)

attracting electrodes at the opposite end of the tube. A fixed scanning aperture is also located at the anode end of the tube and the electrons from the cathode are magnetically deflected by external coils -- as the electrons are moved past the aperture they enter and are amplified by the electron multiplier structure. Therefore, in the image dissector the electronic image is moved while the scanning device is stationary; the opposite is true of the iconoscope.

In 1929 Dr. Zworykin demonstrated a television transmitter based on mechanical scanning with a receiver in which an improved form of cathode-ray tube called the *kinescope*,* was used to reproduce the transmitted 120-line image. In 1931 RCA made experimental television transmissions over station W2XBS in New York City and RCA's president, David L. Sarnoff, predicted that within five years television would become "as much of a part of our life" as radio. As with so many other, similar predictions, however, it proved to be premature.

It was to be four more years before Dr. Zworykin had developed the iconoscope to the point where it could be used as the basis of a workable, all-electronic television system.

After several years of experimentation with mechanical scanning systems, RCA built an entirely new television transmitter at the Empire State Building and equipped NBC's nearby broadcasting studios for broad experimentation in all phases of television broadcasting. In the summer of 1936 RCA began extensive field tests from the Empire State Building with electronically-scanned 343-line pictures, 30 frames per second. In January of the following year, however, definition was raised to 441 lines in accordance with the proposed standards of the Radio Manufacturer's Association, a figure which remained until 1941.

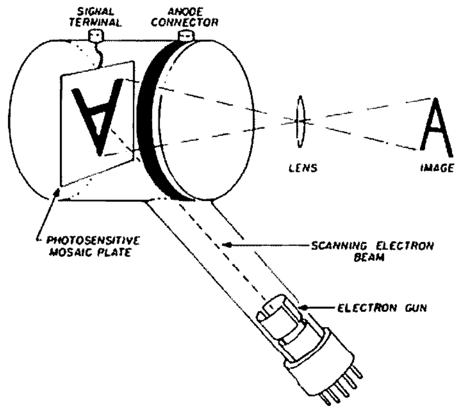


fig. 4. Iconoscope invented by Vladimir Zworykin uses photosensitive mosaic deposited on thin mica plate.

The television art was also advancing in other parts of the world. In England, Electrical and Musical Industries (EMI) set up a TV research group in 1931 under the direction of Isaac Schoenberg. He fostered the evolution of a practical system based on a camera tube known as the Emitron, which was an advanced version of Zworykin's iconoscope, and a CRT for the receiver. Schoenberg saw the need to establish standards that would en-

*From Greek *kine*, "motion," and *scope*, "to observe."

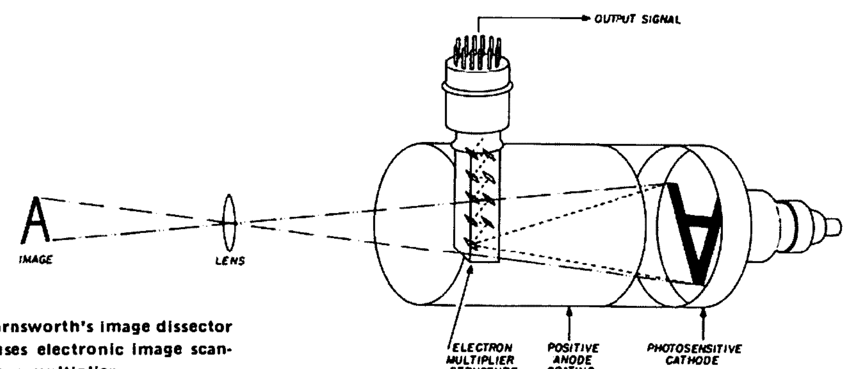


fig. 5. Philo Farnsworth's image dissector camera tube uses electronic image scanning and electron multiplier.

dure for many years and proposed 405-line pictures, 50 frames per second.

The British government authorized the BBC to adopt these standards as well as the complete EMI system, launching the world's first public, high-definition TV service in 1936. These same standards remained in effect until 1964, when they were gradually superseded by a 625-line standard.*

Initially, and for only a short time, the EMI system was under comparison with alternate broadcasts from a 240-line, 25-frame system developed by John Baird. However, the Baird system used mechanical scanning and suffered from poor sensitivity.

electron velocity is required in the neck of the tube, however, because of difficulties in magnetic deflection and focusing with a low-velocity electron beam.

Photoelectrons are emitted from the cathode surface in direct proportion to the light and shade in the scene, converting the optical image into an electron image. The electron image is accelerated toward the target (which is 300 volts positive in respect to the photocathode), and is focused through the screen onto the target plate by a uniform magnetic field in a manner very similar to that used in Farnsworth's image dissector tube. As the electron beams scans the target, a charge distribution corresponding to the picture elements in the light image determines the number of scanning electrons returned to the electron gun.

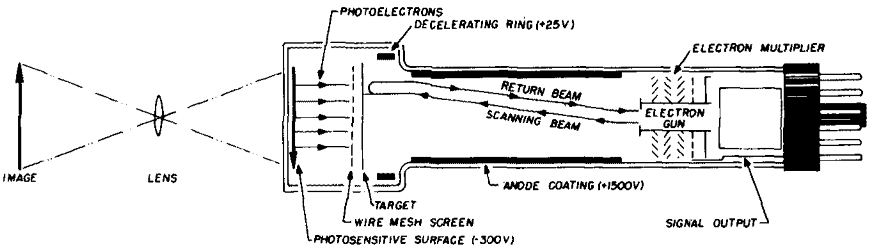


fig. 6. Basic construction of the image orthicon. The external focusing and deflection coils are not shown.

Regular television broadcasts began in Germany in 1935, though with medium definition (180 lines), and in France engineers were working on a high-resolution 1000-line system which eventually resulted in France's 819-line standard.

camera tube development

Later research in television camera tubes resulted in the development of pickup tubes, based on the iconoscope principle, which had greatly increased sensitivity. The first of these was the orthiconoscope or *orthicon* which was developed by Albert Rose and Harley Iams in 1939. Continuing research led to the development of the *image orthicon* by Albert Rose, Paul Weimer and Harold Law of RCA in 1943.

In the image orthicon, **fig. 6**, a glass plate coated on one side with a conducting layer of photoelectric material serves as the photocathode. The semitransparent plate receives the light image on one side while photoelectrons are emitted from the other side, which faces a wire-mesh screen and target, to produce an electron image which corresponds to the scene focused on the front of the glass plate.

The electron gun produces a stream of electrons which is accelerated toward the target by the positively charged anode wall coating. Beam deflection is accomplished with magnetic deflection coils which are mounted externally on the tube. A decelerating ring with a very low positive potential is placed near the target to slow down the electrons so the scanning beam does not have sufficient velocity to produce secondary emission that generates spurious shading signals. High

*There is still one BBC station broadcasting 405-line telecasts. At this time no firm date has been established to convert it to the 625-line standard.

The returning stream of electrons arrives at the gun close to the aperture from which the electron beam emerged. When the returning electrons strike the aperture disc, which covers the gun element and is at a potential of about +200 volts, they produce secondary emission. Therefore, the disc serves as the first stage of a five-stage electron multiplier -- the output current from the final stage varies in magnitude with the light image. A more complete discription of this complex tube is contained in reference 6.

While the image orthicon still plays a dominant role in television broadcasting, the more compact *vidicon*, introduced in the early 1950s, is used in most amateur TV systems. The vidicon, **fig. 7**, makes use of a semiconducting material which is characterized by a resistance that decreases upon exposure to light.[7] The inside surface of the glass faceplate is coated with a very thin layer of photoconductive material; the optical image is focused on the other side of the plate and the photoconductive layer is scanned with an electron beam which deposits just enough electrons on each spot it touches

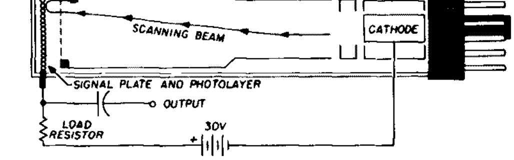


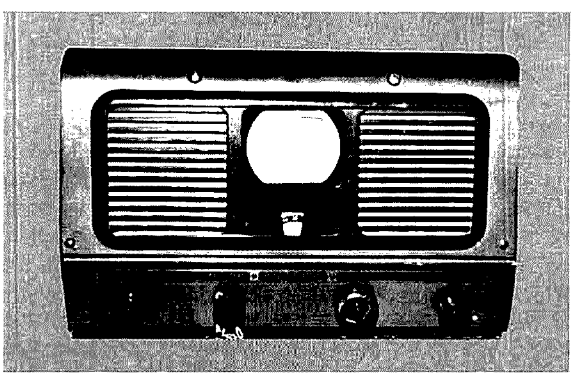
fig. 7. Schematic diagram of the vidicon. External focusing and deflection coils are not shown.

that it reduces the signal plate-to-cathode potential. During the short time between successive scans, charge leaks through the photoconducting material at a rate which is determined by the intensity to which that part of the photoconducting material is subjected. As the electron beam scans the surface of the photoconducting material, the charge it deposits varies in accordance with the variations in the illumination of successive elements of the photoconductor. Therefore, the current through the load resistor, and hence the output voltage, electronically reproduces the light intensity of the scene.

frequency allocations

Late in the fall of 1937 the FCC announced new allocations for the spectrum between 30 and 300 MHz and, much to the delight of amateurs, reaffirmed the 56-60 MHz (5-meter) band as exclusively amateur. The new rules also provided two new exclusive amateur uhf bands: 112-118 MHz (2½ meters) and 224-230 MHz (1¼ meters). One of the big worries at the time was the huge spectrum space demanded by the impending arrival of television, still around several corners but getting closer. In fact, the Commission's press release on the new uhf allocations commented that, "The investigations and determinations of the Commission justify the statement that there does not appear to be an immediate outlook for the recognition of television service on a commercial basis. The Commission believes that the general public is entitled to this information for its own protection . . . "

Nevertheless, the FCC allocated seven main television channels, each 6 MHz wide, between 44 and 108 MHz, and twelve additional channels above 156 MHz. The 50-56 MHz TV channel was of special concern because of possible interference due to its close proximity to the amateur 5-meter band. In New York this channel was assigned to CBS, and in a brief survey their engineers logged scores of amateur stations operating between 54 and 56 MHz, well outside the band. When you consider that modulated oscillators and superregen receivers were the order of the day, this is understandable, but the new TV allocations spelled the end of broad signals from unstable 5-meter transmitters (which were often operated on raw ac). Not unexpectedly, in December, 1938, the FCC required that all 5-meter amateur transmitters meet the same stability requirements as those already imposed on the lower frequencies.



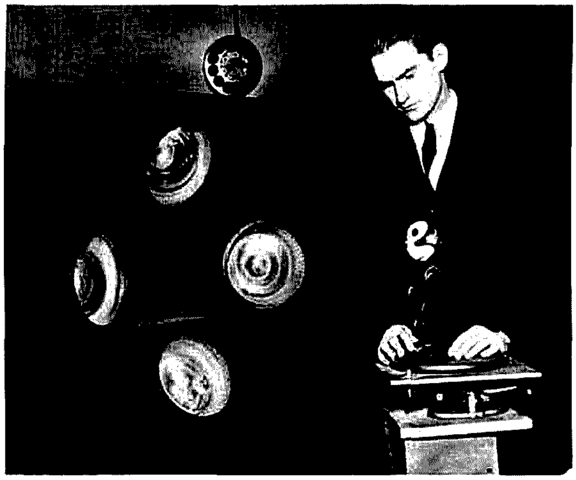
Pilot home television set from the late 1930s, one of the first sets offered to the consumer. From the AWA Museum collection. (Photo by W2BWK)

This television set, first introduced by RCA for public use at the New York World's Fair in 1939, featured a picture reflected from the top of the kinescope to a mirror on the underside of the cabinet's uplifted lid. (Photo courtesy RCA)

modern television

The first regular television schedule in the United States was introduced by NBC's W2XBS in 1939 with a telecast of President Roosevelt opening the World's Fair in New York. RCA announced the new NBC programming in an advertisement for television receivers in *QST* which explained that NBC stations in New York, Schenectady and Los Angeles would begin telecasting two one-hour programs per week, plus special pickups of sports, visiting celebrities, etc.[8] The public, however, didn't respond eagerly to the new medium, and after five months of broadcasting, RCA had sold only 400 television sets. The story was much the same in England where only 3000 receivers had been sold after two years of television broadcasting by the BBC.

The New York World's Fair also marked an important milestone for amateur television. The Managing Director of W2USA at the World's Fair, Art Lynch, W2DKJ (now W4DKJ), after seeing a successful demonstration of amateur television equipment at a radio show in Chicago in June, was convinced that television communications should be added to the station at W2USA, "the most visited amateur station in the world." Since the World's Fair was scheduled to close at the end of October, time was short, but Art lined up the necessary talent, and with some help from industry, the group built two complete television systems in an effort to establish the first *two-way* television contact. Their goal was accomplished on September 27, 1940, when amateurs at W2USA and W2DKJ/2 at the New York Daily News Building in Manhattan began exchanging fair quality television pictures



In the early 1930s, Felix the Cat was the first "star" to appear before RCA-NBC experimental television cameras. Felix whirled around on his phonograph turntable for hours on end while four hot arc lights beat down on him. In those early days the crude TV images of Felix looked like he was being viewed through a venetian blind. (Photo courtesy RCA)

on the amateur 112-MHz band.[9] Accompanying sound was transmitted on 56 MHz. Distance between the two stations was about eight miles.

The television equipment at each end of the circuit consisted of a camera-modulator unit, a receiver and a transmitter which were duplicates of equipment described earlier in *QST*.[10-12] The system used 30-Hz vertical scanning, 3600-Hz horizontal scanning and a 120-line raster. Considering that the pictures were viewed on a CRT with a P1 phosphor, the results were quite gratifying. Each station boasted the very latest in electronic equipment including electro-magnetically-deflected cathode-ray tubes, free-running sweep circuits synced by external pulses and iconoscope camera tubes. The equipment was donated by RCA, National, Hallicrafters, Hammarlund, Thordarson and Kenyon. The station at W2USA used a single 1000-watt lamp for subject illumination while W2DKJ/2 had a battery of smaller lights with reflectors.

A number of amateurs in the vicinity of New York were working on their own television receivers and on October 15th, W2AOE put on a demonstration for members of the Northern Nassau Radio Association by receiving TV signals from the 20-watt station at W2DKJ/2, 17 miles away, using an improved version of the receiver described by J. B. Sherman in *QST*.[9] The range was increased to 29 miles on October 19th when good quality TV signals from W2DKJ/2 were received at W3FRE in Denville, New Jersey.

On July 1st, 1941, NBC's New York station, called WNBT, and CBS's station, WCBW, were licensed as the first commercial television stations in the United States. The FCC authorization provided for an upgrading in picture definition by adopting a 525-line standard, and fm for the audio portion of the telecasts (replacing a-m). However, the outbreak of the war in December brought television broadcasting to a standstill, and as critical materials and manpower were channeled into the war effort, television broadcasting ceased.

The FCC was carefully studying spectrum allocations during the last few years of the war, in anticipation of the armistice, and in March, 1945, they announced the new vhf allocations above 108 MHz and below 44 MHz. The spectrum between 44 and 108 MHz was to be allocated later, after running fm transmission tests during the summer. Since the release of raw materials was not imminent, this didn't appear to pose any problem. However, after VE day cutbacks and labor layoffs commenced in industry and it appeared that needed raw materials would soon be available — on June 27th the FCC announced the allocations between 44 and 108 MHz without running their planned tests. Under the new plan amateurs would get 50-54 and 144-148 MHz, fm broadcasting would move to 88-106 MHz (106-108 MHz was reserved for facsimile broadcasting), and television received channels 1 through 13. Channel 1, originally slated for the 44-50 MHz slot, was later deleted.

By 1948 there were 36 television stations on the air, 70 more were under construction, an estimated one-million television sets were in use by the public, and interference problems began to appear. In September, 1948, the FCC put a freeze on licensing any new TV stations in order to study the frequency allocations and to consider the problems posed by color television (more about that later). This situation continued for three years, prolonged by the Korean War and a consequent shortage of critical materials. Finally, in April, 1952, the FCC lifted the freeze with a document that supplemented the twelve existing vhf channels with 70 new uhf

Here's how Felix the Cat looked on the screens of experimental black-and-white TV sets in the early 1930s. The picture was transmitted by RCA-NBC cameras from a studio in New York City, and was received as far away as Kansas. There, and at points in between, it was picked up by video bufts on their primitive 60-line viewers. (Photo courtesy RCA)

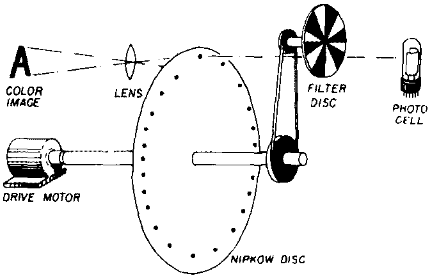


fig. 8. Color television system of the late 1920s. Light from the image is concentrated by lenses on the main scanning disc, but reaches photocell only when the proper color filter is presented by the second rotating disc, which revolves faster than the scanner. Similar setup was used at the receiver.

channels. Within a few months they had processed a backlog of 700 applications for new stations and had granted 175 new licenses. Within a year there were 377 stations on the air, and by 1955 about 95 per cent of the country had television coverage. Today there are 919 television stations (590 on vhf, 329 on uhf) throughout the United States and there are few places in the world that don't have television service.

color television

Although color television is generally accepted as a product of the past 25 years, it is nearly as old as television itself. One of the earliest proposals was patented in Germany in 1904, and the same Dr. Zworykin who invented the iconoscope filed a patent disclosure for an electronic color TV system in 1925.

John Baird demonstrated the first practical color television system in 1928 which used a Nipkow disc with three spirals of 30 apertures, one spiral for each primary color. The light source at the receiver used two gas-discharge tubes: one of mercury vapor and helium for the green and blue colors, and a neon tube for red.

In 1929 Herbert Ives and his colleagues at Bell Laboratories transmitted 50-line color images between New York and Washington, DC. This was also a mechanical system, but one that simultaneously sent the three primary color signals over three separate circuits.

In 1940 both NBC and CBS gave public demonstrations of color television which used 441-line scanning. Numerous demonstrations were also given after the war, including one by RCA in 1946 in which a stereoscopic system was used to present a three-dimensional representation of the image. In all of these demonstrations, however, color filter discs (or drums) rotated in synchronism in front of the camera tube and receiver.

At the receiver the color images were presented sequentially (field-sequential system) so the red, green and blue components of the scene were viewed one after the other. Because of the persistance of vision the viewer perceived a full-color image; however, if he moved his head or scanned the picture rapidly, the image suffered from "color break-up." The rotating mechanical discs were also a drawback, and as black-and-white TV sets became widely distributed in the late 1940s, the inability of unmodified monochrome receivers to reproduce a color program made color television broadcasting, on this basis, economically impractical.

These difficulties were solved by a simultaneous three-channel color system introduced by RCA in 1946 in which the three component images (red, green and blue) were separately transmitted and projected on a screen or presented on three separate CRTs which were viewed through a system of beam-splitting dichroic mirrors. RCA even developed a projection CRT for this purpose which they called the *trinoscope*. Monochrome receivers were simply tuned to the green channel (fig. 9).

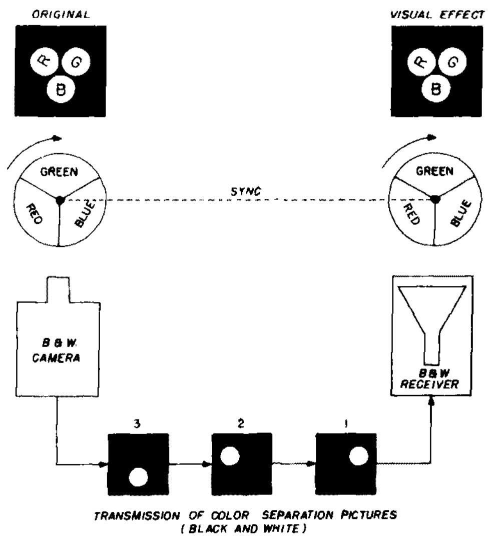


fig. 10. Basic system for field-sequential transmission of color television images.

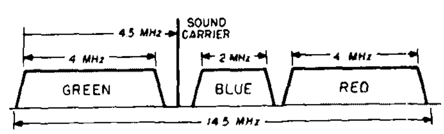
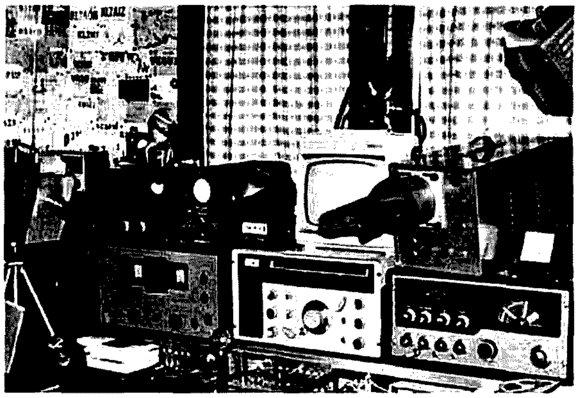


fig. 9. Transmission channel for RCA's experimental simultaneous color television picture signal required 14.5-MHz bandwidth. Monochrome receivers were tuned to the green carrier. Detail capable of being resolved in a blue image is much less than in a green, red or white image so bandwidth of blue video signal can be reduced substantially without affecting the quality of the color picture.

However, both the field-sequential and simultaneous three-channel color systems required, for equal picture definition and freedom from flicker, much greater bandwidth than the 6-MHz channels already allocated to black-and-white TV. In view of the great pressure for frequency allocations in the vhf spectrum, it was generally agreed that color television should be accommodated within the existing 6-MHz channels. By reducing both the color frequency and the number of lines, the field-sequential color system could be transmitted within a 6-MHz bandwidth, but only with poor resolution and increased flicker.



Slow-scan setup used by Don Miller, W9NTP, in 1969. Equipment was very up to date at the time, included a sampling slow-scan camera (left), a shuttered camera (upper right) and both slow- and fast-scan monitors. Don's sampling camera was one of the first of its type and was widely duplicated by amateurs.

Investigators at RCA (1949) found it was possible to retain full resolution, freedom from flicker, *and* monochrome compatibility with a simultaneous system that used a monochrome picture signal with a phase and amplitude-modulated subcarrier which carried the color or chroma information. The chromatic subcarrier, approximately 3.58 MHz above the picture carrier, was selected so it had no visible effect on the picture reproduced by a monochrome receiver. In a color receiver the subcarrier was used to distribute picture brightness between the three primary colors to produce a natural color rendition of the original scene.

Nevertheless, in October, 1950, after a lengthy series of hearings, the FCC adopted the incompatible field-sequential color system as the standard for the United States. However, in December, 1953, the Commission rescinded its earlier ruling and issued a new set of specifications which had been submitted by RCA and the National Television System Committee (NTSC). These corresponded to the compatible color system developed earlier by RCA – this same basic color system is still used throughout North and South America, Japan, Korea, and parts of Europe.

slow-scan television

No history of television would be complete without some mention of slow-scan television, and the important role that amateurs played in its development. Copthorne MacDonald, W4ZII (now WØORX), introduced slow-scan television to amateurs in a 1958 *QST* article[13] which described a simple system, using a flying-spot scanner, to transmit photo transparencies. Initial on-the-air tests were conducted on 11-meter a-m between W4JP at the University of Kentucky and K4KYY. MacDonald also tried to run tests with PJ2AO in Curacao, but band conditions were too poor for satisfactory picture reception.

The slow-scan system, which requires no more bandwidth than an audio signal, was originally conceived as a facsimile system and it was a number of years before the medium was used to transmit live images. Since 11 meters was the only high-frequency band where facsimile transmission was permitted, most sstv activity ceased when amateurs lost 11 meters to the Citizens Radio Service. Eventually, however, the FCC granted special permission to conduct sstv tests on 10 meters and, later, 20 meters. The sstv standards which are used today were developed during these early tests. Since August, 1968, slow-scan television (designated narrow-band A5 and F5 emission) has been permitted on portions of all the high-frequency bands plus most of vhf.*

*Complete bibliographies of slow- and fast-scan television articles which have appeared in *QST* are available from ARRL, 225 Main Street, Newington, Connecticut 06111. Send a stamped, self-addressed, business-size envelope.

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ham radio

the 1979 World Administrative Radio Conference

and what it means to you

What, me worry? Yes, you worry – about an increasingly important four-letter word – WARC.

What's a WARC? A WARC is a World Administrative Radio Conference – a gathering of all the ITU member nations to examine and decide upon basic questions of mutual interest. The first WARC was in Berlin in 1903, the most recent in 1959, and the next has been scheduled for sometime in the second half of 1979, in Geneva, Switzerland.

Okay, that's a WARC – so what? The "What" of the WARC is *what* is going to be discussed in 1979 – namely, frequency allocation, or perhaps more properly, *re*-allocation, that's "what."

On an international basis the radio spectrum from 10 kHz to 47 GHz has already been allocated. There are no unallocated segments of the spectrum *within* those limits. Therefore, if some user of that part of the spectrum needs additional frequencies, it must come from someone else's *present* allocation (*ah* – the light dawns!). Yes, even from the hitherto sacrosanct domain of the amateur bands if the justification for such a request is strong enough. And there is the secret word – justification! Say it correctly, and the duck will bring you 200 kHz (sorry, Groucho).

Seriously, though, justification is not the numbers game – just the number of licensees alone in a given radio service will not be adequate justification for getting new frequencies, much less keeping those already allocated. Sure, the Amateur Radio Service has grown from 46,000 in 1934, to 185,000 in 1959, to 275,000 today, with basically the same allocated spectrum – give or take a hundred kHz. Crowded? Sure. QRM – you bet! But can you imagine what it would be like if we still used only double-sideband-with carrier? Absolute chaos!

The Amateur Radio Service responded to increased band crowding in its historical manner – ingenious adaptations of, and subsequent improvements upon, commercial techniques to relieve congestion. (Sounds like a nasal spray commercial but it sure worked – remember the disparaging remarks about the "Donald Duckers," and when SSB was known as SSSC?)

Well, if numbers aren't the answer, then what is? Simple, and like many other things, money included, it's not how much you have, but what you do with what you have. How *does* Amateur Radio use its allocated spectrum? Is it being used wisely for the benefit of the public at large and in keeping with the Service's Basis and Purpose as outlined in Part 97.1 of the FCC Rocks and Shoals? Or, is it being used for the personal amusement and satisfaction of a miniscule percentage of the U.S. citizenry? What are the trends in the Amateur Radio Service? Where will it be in the year 2000? How will, or can, WARC influence this?

These and similar questions, plus those dealing with the Amateur frequency needs now and up to the year 2000, are being discussed by members of eight task forces set up by the FCC in what's called the "Amateur Working Group." This group, numbering about forty, has been given the job of developing recommendations for the United States Amateur Radio Service position in the next WARC – including the justifications needed to keep the frequencies it presently has, plus – maybe – getting some new ones. There's plenty of time 'till 1979, right? *Wrong!*

The lead-time of a bureaucratic, international operation like a WARC boggles the mind! What with the need to coordinate, review, correlate, adjust, modify, etc., both within the FCC and between various parts of our government, coordinate *unofficially with other* ITU member governments, and so forth, it's not surprising that the preliminary Amateur Radio Service frequency allocation request has to be in the FCC's hands by the time you read this! And there are only a few more months in which to come up with the most persuasive justification possible for the continuation of the Amateur Radio Service as we now know it, and would like it to be. This "Amateur Radio Service position paper," as it is being called, has to be submitted to the FCC for its consideration no later than June of this year – *this* year, not 1979!

In the meantime, if you're now concerned, re-read Stu Cowan's excellent article in the April, 1965, issue of *QST*.*

What, me worry . . . you bet!

By Pete Hoover, W6APW, 1520 Circle Drive, San Marino, California 91108

*Stuart D. Cowan, W1RST, "The Death – or Survival – of Amateur Radio," *QST*, April, 1965, page 80.

MICROPROCESSORS

What is a microcomputer input/output device?

In the discussion of the anatomy of a microcomputer last month, we described the various data paths in a microcomputer, including data input, data output, external device addressing, in and out function pulses, and interrupt signals. These are the vital lines of communication between the microcomputer and the "outside world," i.e., those signal lines that are necessary to interface the microprocessing unit (MPU) to the input/output, or I/O devices that you would like to control.

What, exactly, is an I/O device? Some useful definitions include the following:

input/output General term for the equipment used to communicate with a computer and the data involved in the communication.[1]

I/O Abbreviation for input-output.[2]

I/O device Input/output device. Any digital device, including a single integrated-circuit, that transmits data to or receives data or strobe pulses from a computer. The in and out functions are always referenced to the computer.[3]

The traditional view of an I/O device is that it is somewhat large or complex. Card readers, magnetic tape units, CRT displays, and teleprinters certainly fit such a description. However, a single integrated-circuit chip, such as a latch, shift register, counter, or small memory can also be considered to be an I/O device to a computer.

Another important point is that several device-select pulses may be required to interface a single I/O device. For example, a 74198 shift register has a pair of control inputs that determine whether the register shifts left, shifts right, or parallel loads eight bits of data. This chip also has a clock input and a clear input. Thus, a single 74198 chip, when serving as an output device, may require up to four device-select lines from the microcomputer. Therefore, the fact that we can generate 256 different input and 256 different output device select pulses does not necessarily mean that we can address 512 different "devices." A more reasonable number is of the order of 50 to 100 different devices.

Device-select pulses are inexpensive and easy to implement. We encourage you to use them as often as possible as you attempt to substitute computer software, (microcomputer programs) for integrated-circuit chip hardware. We shall repeat this theme often: software vs hardware. There is a tradeoff between the two, but your main objective in using microcomputers will usually be to substitute software for hardware. When you do so, the only penalty that you may pay is time because it takes time to execute computer instructions. If you can accept the delays inherent in computer programs, then you can vastly simplify the circuitry required to accomplish a specific interfacing task.

By Peter R. Rony, David G. Larsen, WB4HYJ, and Johathan A. Titus.

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

what is interfacing?

Interfacing can be defined as the joining of members of a group (such as people, instruments, etc.) in such a way that they are able to function in a compatible and coordinated fashion.* By "compatible and coordinated fashion," we usually mean synchronized. Some important definitions include the following:

Synchronous — In step or in phase, as applied to two devices or machines. A term applied to a computer in which the performance of a sequence of operations is controlled by equally spaced clock signals or pulses.[2] At the same time.

Synchronous computer — A digital computer in which all ordinary operations are controlled by equally spaced signals from a master clock.[2]

Synchronous operation — Operation of a system under the control of clock pulses.[3]

To synchronize — To lock one element of a system into step with another.[2]

Synchronization pulses — Pulses originated by the transmitting equipment and introduced into the receiving equipment to keep the equipment at both locations operating in step.[2]

We can thus define *computer interfacing* as "The synchronization of digital data transmission between a computer and one or more external input/output devices."[3]

Although the details of computer interfacing vary with the type of computer employed, the general principles of interfacing apply to a wide variety of computers. Such principles include the following:

The digital data that are transmitted between a computer and an I/O device are either individual clock pulses or else full data words.

The computer and the input/output device are both *clocked* devices. At the very least the I/O device has a single flip-flop that is set or reset by the computer. All data transmission operations are synchronized to the internal clock of the computer.

The computer sends synchronization pulses, called device-select pulses, to the I/O device. These pulses are generated by the computer program i.e., they are software generated, and are usually quite short (for an 8080 microcomputer operating at 2 MHz, they last only 500 nsec). *They synchronize and select at the same instant of time.*

Individual device-select pulses are sent to individual input or output devices. This is called *external device addressing*. The pulses are used for latching data output and strobing data input.

Computer program operation can be interrupted by the transmission of a clock pulse from an I/O device to a special input line to the computer. This is called *interrupt generation*. Upon being interrupted by an external I/O device, the computer goes to a computer subroutine that responds to, or *services*, the interrupt.

Full data words can be output from, or input into, the accumulator register. For the 8080 microcomputer, a full data word contains eight bits. Output data from the accumulator is available for only a very short period of time, and usually must be latched. Input data into the accumulator is acquired over a very short period of time, and usually must be strobed into the accumulator.

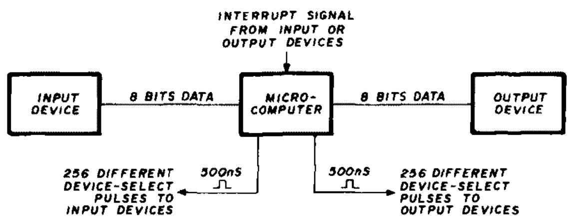

fig. 1. The I/O and control data paths in an 8080 microcomputer.

As shown in fig. 1, which summarizes the above comments, interfacing basically consists of the synchronization of parallel input or output data via the use of the 512 device-select pulses.

Hardware is required to tie the MPU to the external device and is just as important as the microcomputer software. We shall tackle both of these facets of microcomputer interfacing in detail in subsequent columns. In the next column we will discuss the output instruction for the 8080 microprocessor chip, which has at least four sources of supply. This is more than for any other MPU. It is clear that the 8080 MPU is destined to become a widely used microprocessor.

*Charles L. Garfinkel of Keithley Instruments, Inc. is the originator of this definition.

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ham radio

horizontal-antenna gain

at selected vertical radiation angles

Graphical aids for choosing antenna height to optimize near- or far-field gain

In a previous article, I gave the gain for certain vertical-plane radiation angles for horizontal antennas at certain antenna heights.[1] Since then I've had requests for data on what height to use for optimizing gain at a certain radiation angle. Suppose you'd like to work DX on 20 meters and your beam is on a 40-foot (12m) tower. You may find in this case that you hear stations about 900 miles (1440km) away much louder than the DX stations, which are perhaps over 2000 miles (3200km) away. Would it help, and if so, how much, to use a higher tower? The answer to these questions are in this article. The data is useful in selecting heights for horizontal dipoles for 40 and 80 meters as well as tower heights for beams at 10, 15 and 20 meters. Most of this article and examples, however, cover the latter case.

antenna height for several radiation angles

The answers to the question of what height to use to optimize gain at certain radiation angles is given in **fig. 1**, in which relative antenna gain is on the vertical scale and antenna height in wavelengths is on the horizontal scale. Curves are given for several radiation angles, α. Reference gain is 1.0 (the gain of a half-wave antenna in free space for all radiation angles). This is just a convenience, as all the gains are relative. If an antenna is higher than a half-wavelength, multiple lobes occur in the vertical plane; this data is shown in detail in the ARRL *Antenna Handbook*.[2] Similar data is also shown in **fig. 1** especially for $\alpha = 30°$, which shows peaks near $h/\lambda = 0.5$ and 1.5, with nulls at $h/\lambda = 1.0$ and 2.0.

Fig. 1 shows that for $\alpha = 5°$, the higher the tower the better, as the peak in the gain curve doesn't occur until the antenna height is almost three wavelengths ($h/\lambda = 3.0$). As a convenience **table 1** is included, which gives h/λ for tower height for the 10-, 15- and 20-meter bands. Thus, only at 10 meters with a 100-foot (30m) tower is $h/\lambda = 3.0$, where maximum gain is achieved at $\alpha = 5°$. **Table 2** shows the relationship between radiation angle, α, and distance for F_2-layer one-hop signals. **Fig. 1** also shows that for radiation straight up ($\alpha = 90°$), a very low antenna ($h/\lambda = 0.1$) is sufficient; for $\alpha = 10°$, an $h/\lambda = 1.2$ is best; and for $\alpha = 15°$, a first plateau in gain is reached at $h/\lambda = 0.6$, with maximum gain at $h/\lambda = 1.1$.

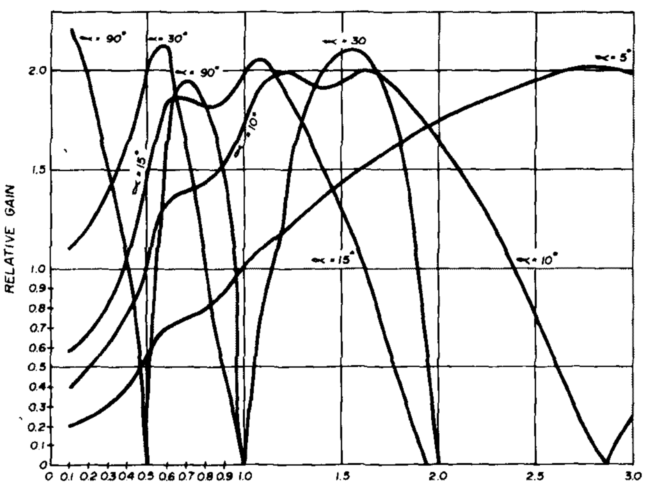

fig. 1. Horizontal-antenna gain as a function of height, h/λ, for several vertical-plane radiation angles, α

The graph would become pretty messy if more α values were plotted. (I have data for other α values, which I'd be happy to send on request.) The curves shown should cover most situations since there's not as much need to optimize antennas for vertical-plane radiation angles above 15° or for distances less than 1200 miles (1920 km) as there is for DX work.

other examples

It's also useful to plot gain versus tower height, either

By Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University.

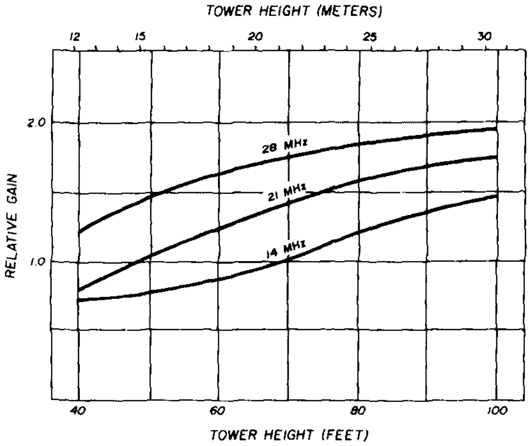

fig. 2. Horizontal-antenna gain as a function of height for the 10-, 15-, and 20-meter bands, $\alpha = 5°$.

for one radiation angle or for one band (**figs.** 2 and 3). **Fig.** 2 shows that for $\alpha = 5°$ (2000 mile or 3200 km one-hop F_2-layer DX), the higher the tower the better for all three bands (10, 15 and 20). Now consider **fig.** 3.

table 1. Horizontal-antenna height in terms of wavelength for the 10, 15 and 20-meter bands.

height		height, wavelengths		
feet	(meters)	10	15	20
10	(3)	0.3	0.20	0.15
20	(6)	0.6	0.40	0.30
30	(9)	1.0	0.6	0.45
40	(12)	1.2	0.8	0.60
50	(15)	1.5	1.0	0.75
60	(18)	1.8	1.2	0.9
70	(21)	2.1	1.4	1.1
80	(24)	2.4	1.6	1.2
90	(27)	2.7	1.8	1.4
100	(30)	3.0	2.0	1.5

If you have a 40-foot (12m) tower, the gain for DX signals ($\alpha = 5°$) is about 0.75, while for signals 900 miles (1440 km) away ($\alpha = 25°$), the gain is 2.25. This ratio is 0.75/2.25 or 3 to 1 against you. This is why those W9s sound so loud and the DX so weak!

If you had an 80-foot (24m) tower, the DX/W9 ratio at $\alpha = 5°$, say, would be 1.26/0.16 = 7.88, which is a lot better (24 times better) than with the 40-foot (12m) tower. The whole business is, however, not too simple as even at 80 feet (24m), the gain for 1200-1500-mile (1920-2400 km) ($\alpha = 15°$ to $10°$) signals is much greater than for 2000-mile (3200 km) distant DX signals; and if you go to a 100-foot (30m) tower, the 900-mile (1440 km) signals are strong and the 1000-mile (1600 km) signals weak! Data is given in **table 3** so that you can plot graphs similar to **figs.** 2 and 3 for other angles or other bands.

table 2. Vertical-plane radiation angle for horizontal antennas for various distances using F2 layer, one-hop propagation.

radiation angle, α (degrees)	distance miles	(kilometers)
2	2600	(4160)
5	2000	(3200)
10	1500	(2400)
15	1200	(1920)
20	1000	(1600)
25	900	(1440)

table 3. Horizontal-antenna gain, height and vertical-plane radiation angle for the 10-, 15- and 20-meter bands.

	40-foot (12m) tower			80-foot (24m) tower		
α	G_{10}	G_{15}	G_{20}	G_{10}	G_{15}	G_{20}
5	1.22	0.80	0.72	1.85	1.6	1.22
10	2.0	1.45	1.36	0.96	2.0	2.0
15	1.9	1.83	1.86	1.32	1.05	1.9
20	1.11	1.88	2.15	1.75	0.65	1.11
25	0.13	1.62	2.25	0.10	1.90	0.13

	60-foot (18m) tower			100-foot (30m) tower		
α	G_{10}	G_{15}	G_{20}	G_{10}	G_{15}	G_{20}
5	1.65	1.22	0.88	1.98	1.78	1.45
10	1.80	2.0	1.57	0.26	1.64	1.98
15	0.40	1.9	1.86	1.97	0.21	1.29
20	1.28	1.11	1.74	0.34	1.84	0.16
25	1.95	0.13	1.28	1.96	1.66	1.49

While there's no simple answer of what tower height to use, it's evident that one at 80 feet (24m) is much more desirable than one at 40 feet (12m) for 20-meter operation to improve the DX/W9 signal ratio. Graphs such as that in **fig.** 3 for 10 and 15 meters are even more complex, so practical and economic factors may dictate which height to use.

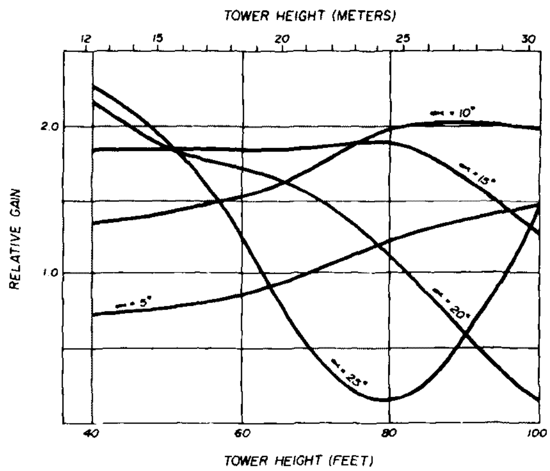

fig. 3. Horizontal-antenna gain as a function of height for the 20-meter band, α = 5, 10, 15, 20, and 25 degrees.

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ham radio

the UAR/T

and how it works

Versatility plus on a 40-pin TTL-compatible chip — useful for many data transmission and receiving applications

One of the largest LSI devices found in recent construction projects and commercial data communications equipment is the UAR/T or universal asynchronous receiver/transmitter. It is also one of the most interesting and versatile chips now available, yet very few people understand its operation. The UAR/T receives and transmits digital information. It acts as a pair of shift registers, the transmitter converting parallel input data to serial output data and the receiver converting serial data bits back to a parallel word. We could easily use an SN74165 as the transmitter (parallel-to-serial) and an SN74164 as the receiver (serial-to-parallel). Data present at the SN74165 is serialized and transmitted to the SN74164 where it is reconstructed again in parallel form.

You could actually perform this experiment, but you would quickly find that a common clock is needed for both shift registers, and the receiver must be synchronized to receive the data as you start to transmit it. If a large number of digital words are being sent between the shift registers, you must have some way to distinguish the end of one word and the start of the next. This requires a great deal of extra synchronizing logic and control lines between the two shift registers.

You probably know that some tricks are used in data communication between terminals and computers, since the data generally flows over one or two pairs of wires and no additional connections are available for clocks or logic control. One of the tricks, using the UAR/T, is to start each data word, generally eight binary bits long, with a START bit and to end each data word with two STOP bits, as shown in **fig. 1**. Now, whenever the receiver is waiting for a new word and it senses the negative edge of the START bit, it resets itself internally and starts shifting in the serial word.

When the two STOP bits are sensed, the data word transfer is complete, and the reconstructed paralleled data is available. Since there are no common clocks, the receiver and transmitter operate out of sync, or asynchronously. The clocks at both ends of the transmission circuit are set very closely but are not exact. The UAR/T makes up for this by sensing input data in the middle of each bit position. If the bits are not exactly aligned they are still sensed correctly, somewhere close to the middle of each bit, as shown in **fig. 1**.

The clock, supplied externally to the UAR/T by a crystal or R/C TTL oscillator, is set at a frequency 16 times the desired bit rate, which allows the internal logic to perform control and sensing functions. Clock inputs for the receiver and transmitter sections of the UAR/T chip are independent and may be set at different bit rates if needed.

By Jonathan A. Titus, Tychon, Incorporated, P.O. Box 242, Blacksburg, Virginia 24060

functional description

A block diagram of the UAR/T is shown in fig. 2. This 40-pin chip has many functions that control the sending and receiving of data and allow programming the UAR/T for certain functions. The number of data bits per word is programmed from five to eight, the number of STOP bits may be selected as one or two, and odd or even parity may be selected or parity may be eliminated from the data word. Five input lines allow the user to program the format of the data sent and received by the UAR/T (**table 1**). The receiver and transmitter are programmed at the same time, so the format of transmitted and received data must be the same. For convenience the *Control Strobe* signal may be left at logic 1 rather than being strobed, which assures that the programming information is always input. In the following examples, pins **34-39** of the transmitter control are programmed at logic 1 giving an eight-bit data word, no parity, and two STOP bits. Active signals are followed by their abbreviation and pin number.

Eight bits of parallel data are entered on the eight transmitter input lines. It is important to note that this

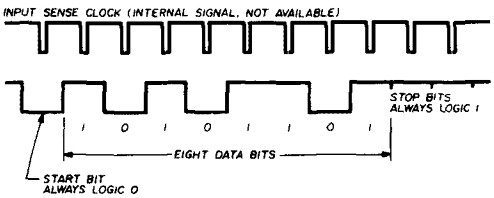


fig. 1. UAR/T data transmission and bit sensing timing.

data may have any format: two BCD digits, an ASCII character, or any random data. Once the eight bits are ready, the $\overline{\text{Data Strobe}}$ ($\overline{\text{DS}}$/23) is pulsed with a logic 0, and the data is transmitted in serial form. The *Serial Output* (SO/25) is a TTL-compatible output that is at a logic 1 when no data is being sent. A *Transmitter Buffer Empty* flag (TBMT/22) is available to indicate that the next eight bits of parallel data may be entered to the UAR/T. The UAR/T is double-buffered, having a holding register as well as the transmitter register. This buffering allows the next data word to be entered and stored while the UAR/T is still transmitting the previous word. The stored word is then automatically placed in the transmitter register and sent.

ASCII keyboard input

A typical UAR/T application is shown in fig. 3 in which an ASCII keyboard supplies the data. The transmitter clock is set at a bit transmission rate 16 times the actual output rate. In the keyboard example there are 11 bits since the START and STOP bits must also be counted. A common telecommunication speed is 110 bits per second, or 110 baud. The clock rate must be 16 times this rate or 1760 Hz, which may be supplied from an NE555 oscillator circuit or other source. Although not used in this example, the TBMT output could signal for the next ASCII character. The TBMT output is often

table 1. Input lines for transmitter control.

	control signal	symbol	function		
35	No parity	NP	0 = no parity entered		
			1 = parity entered		
36	STOP bits	SB	0 = stop bit		
			1 = 2 stop bits		
37-38	Bits per word	NB2,NB1	NB2	NB1	data bits
			0	0	5
			0	1	6
			1	0	7
			1	1	8
39	Parity select	PS	0 = odd parity		
			1 = even parity		
34	Control strobe	CS	enters the above control bits to the UAR/T		

used when data is stored in a buffer or computer and you want to send one word right after another to use the data communication lines efficiently. Whenever TBMT goes to logic 1, the next eight-bit data word is entered to the UAR/T buffer register.

The serial output from the UAR/T can go to an fsk generator to store the data on tape, to a modem, or even to another UAR/T. Although you may not have recognized it, fig. 1 represents the transmission of an ASCII 5 or octal 265. (Remember that the least-significant bit, DB1, is sent first, right after the START bit.) The UAR/T receiver section must be programmed to receive data in the same format as it was sent. The receiver acts as your serial-in, parallel-out shift register, reforming the data into a parallel data word. When the receiver senses a negative transition at the edge of a START bit, it resets to receive a new serial data word. The receiver waits eight clock pulses then starts to sample the serial input bits. This initial offset of eight clock pulses positions the sensing pulse in the middle of each serial data bit, which makes up for the asychronous clocks. The clock difference may be about ±5%.

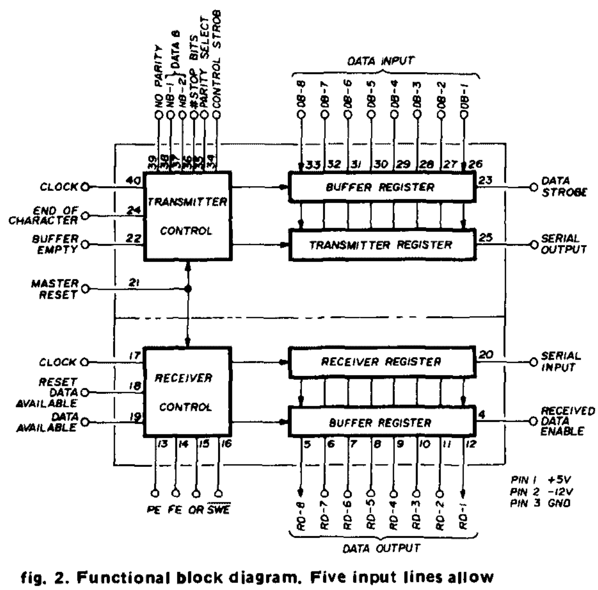


fig. 2. Functional block diagram. Five input lines allow programming of transmitted and received data.

Besides the eight output data lines, the receiver also has some error and flag outputs. The error signals are not frequently used in small systems, but they can serve a useful purpose in debugging systems that use serial data transmission. The *Parity Error* (PE/13) indicates that the parity programmed in the UAR/T and the parity of the received word don't match. The *Framing Error* (FE/14) indicates that the received word doesn't have valid STOP bits, and the *Overrun* (OR/15) indicates that we haven't read the current word and a new word just took its place on the eight output lines. A logic 1 on any of these lines signals an error.

remote data transmission

A *Data Available* flag (DAV/19) goes to a logic 1 to signal that a complete character has been received and may be read at the eight output lines. The data may be read by a terminal (TV typewriter), a computer (Mark-

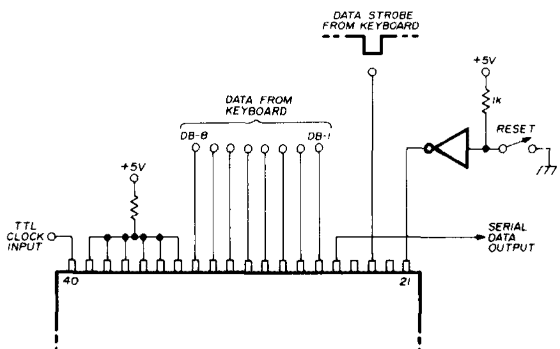


fig. 3. Typical UAR/T application using keyboard input for eight-bit words.

8), or other data storage or output device. After the word is read, the Data Available flag must be reset or it will not indicate when the next word has arrived. Pulsing the Reset Data Available line ($\overline{RDAV}$/18) with a logic 0 resets the flag. If the flag is not reset, the next word received will generate an overrun error.

The receiver's data, error, and flag outputs are all tri-state so that a number of UAR/Ts could be used on a bus input scheme. The Receiver Data Enable ($\overline{RDE}$/4) and the Status Word Enable ($\overline{SWE}$/16) enable the data and flag outputs so that we can read the data. For general, non-bus applications, both these enable lines may be connected to ground. If the UAR/T is to be used on an input bus to a computer or terminal, the tri-state outputs are enabled at the correct time by pulsing $\overline{RDE}$ and $\overline{SWE}$ with logic zeros. In the Mark-8 this is done with input instructions.[1]

Fig. 4 shows how a UAR/T could be connected to the TV typewriter to provide the ASCII input from a remote location, possibly from the keyboard shown in **fig. 3.** In this example, the data-available flag triggers an SN74121 monostable to provide the key-pressed pulse to the TV typewriter, and this pulse is also used to clear the data available flag.

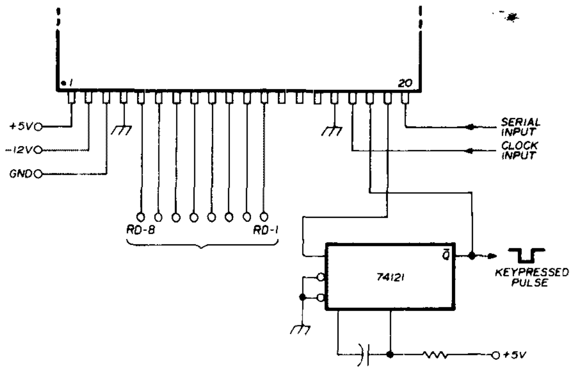


fig. 4. Remote UAR/T receiver for display terminal.

remote data acquisition

Having a receiver and transmitter available in a standard 40-pin package represents a considerable package count, cost, and power saving over a discrete or SSI (small-scale integration) version of this circuit. UAR/Ts have many other applications besides transmitting data back and forth to terminals and computers, so they become useful tools for remote data acquisition and remote control. For example, BCD data could be stored temporarily in a shift register then shifted, one BCD character at a time, to the UAR/T to be transmitted to a terminal or printer. By connecting DB5, DB6, and DB8 to logic 1 and DB7 to ground, octal 260 is inserted into the transmitted data, converting it directly to ASCII. Decimal 3 becomes 263, the ASCII code for 3. The source of the BCD data could be a digital meter, pressure indicator or position encoder -- multiple digits are sent over a pair of wires!

The acquisition and transmission of the data can be controlled by using the receiver section and two SN7485 digital comparators (**fig. 5**). You can compare an output character from the receiver to a preset eight-bit data word. When the two are equal, a monostable starts the data acquisition/transmission sequence and resets the data-available flag. Using a dozen or so 7400-series chips

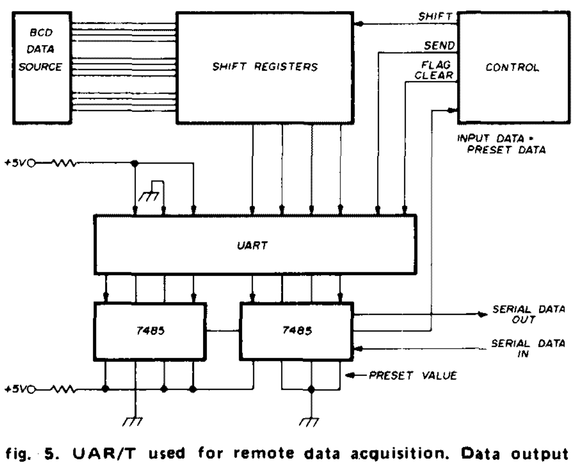


fig. 5. UAR/T used for remote data acquisition. Data output starts when the 7485s find an input equal to the preset value.

and a UAR/T, you now have a four-wire remote data station.

The UAR/T data inputs don't have to be limited to encoded data. They can also be used to monitor limit switches on equipment or even burglar alarm switches or fire sensors. **Fig. 6** shows how two UAR/Ts can be used to indicate remote switch positions. Open and closed switches enter logic 1s or zeros to the UAR/T, and this data lights the corresponding LEDs at the receiver. Data is continuously transmitted by deriving the $\overline{DS}$ pulse from the clock input.

The remote UAR/T receiver section can also be used for housekeeping control at the area being checked. Two SN74154 decoders are connected to the receiver output lines. You can now decode up to 256 possible combinations; and using some NOR gates, you can generate a positive output for each combination. Only one of the 256 combinations may be active at any time. You could also have used eight relay drivers connected to the eight receiver output lines, which would allow independent control of eight devices. The keyboard located at the monitoring station is used to control the receiver outputs. Complete connections in these examples have not been shown for clarity.

The serial output from the UAR/T should not be used to drive lines longer than about six feet (1.8m). If longer lines are required, line drivers and receivers such as the DM8820 and DM8830 should be used. Each of the UAR/T outputs has a TTL fan-out of one load; and although the UAR/T is a mos device, it doesn't require pull-up or pull-down resistors.

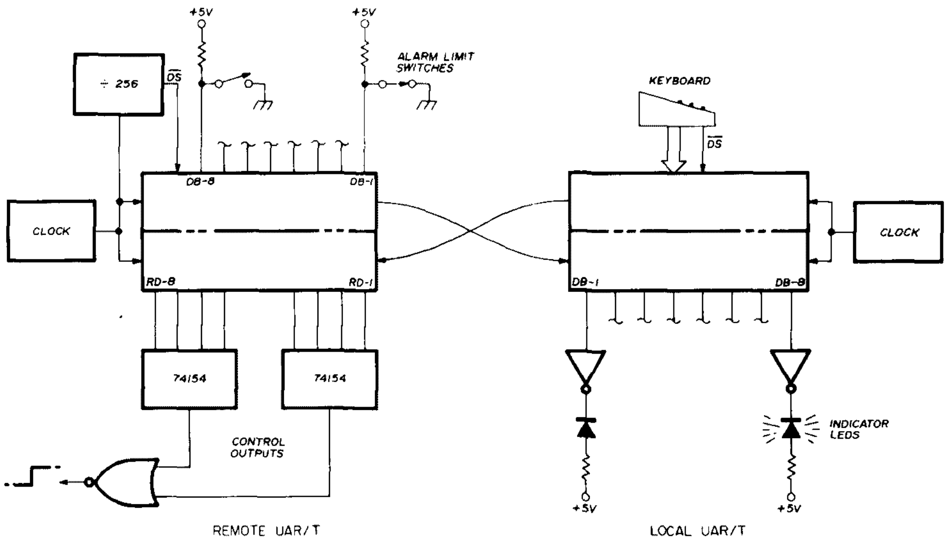


fig. 6. UAR/T used for remote sensing of eight inputs and for remote control.

availability

If you want to experiment with UAR/Ts for data communication or remote control, you may find it difficult to insert the 40-pin chips in breadboard sockets such as those available from E&L Instruments, Continental Specialties, and AP, Inc. To make UAR/T experimenting easy, a special breadboard* has been developed that brings all the connections to 16-pin IC sockets for easy connections with jumpers, and the most important connections are brought to the front of the breadboard to small pins. The complete breadboard plugs into an E&L Instruments SK-10 socket or an AP, Inc. Superstrip socket, leaving plenty of extra room for other chips and connections. Pins on the UAR/T board pick up 5 volts and ground from the power buses. The −12 volts must be supplied with a jumper. All connections are labeled by function and pin number. UAR/Ts available from various manufacturers are generally pin-for-pin compatible, but data sheets should be thoroughly checked before use. The UAR/Ts listed below are compatible.

source	part no.
General Instruments, Inc. 600 West John Street Hicksville, New York 11802	AY-5-1012
Western Digital Corp. 3128 Red Hill Avenue Newport Beach, California 92663	TR1602A & TR1402A
Texas Instruments, Inc. P. O. Box 5012 Dallas, Texas 75222	TMS-6011-NC
American Microsystems, Inc. 3800 Homestead Road Santa Clara, California 95051	S-1883

reference

1. Jonathon Titus, "Computer!" *Radio Electronics*, July, 1974, page 29; "Computer Modifications," *Radio-Electronics*, December, 1974, page 43.

ham radio

*The UAR/T breadboard is available from E&L Instruments, 61 First Street, Derby, Connecticut 06418 as part no. LR-21.

repair bench

Michael James

voltage troubleshooting

Most successful electronic technicians combine a number of different troubleshooting techniques when tracking down a circuit problem, including signal tracing, resistance measurements and oscilloscope checks, but voltage measurements are probably the most popular. They go hand in hand with resistance measurements so you can't understand one without understanding the other, but since resistance measurements are often the next logical step after detecting an incorrect voltage, troubleshooting with an ohmmeter will be discussed next month.

Although voltage troubleshooting is probably the best known, it isn't always the best choice – logic dictates that you should first isolate the problem to one section or stage in the equipment. If your test equipment is limited to a voltmeter you can use voltage measurements to pinpoint a problem area, but other techniques are usually faster. Nevertheless, once you know which circuit to look in, voltage troubleshooting is a quick way of finding the faulty part.

Most modern instruction books and schematics include voltages at each transistor or IC terminal, and some include dc voltages and signal levels at various points in the circuit. To troubleshoot the circuit you start by measuring each dc voltage in the suspected circuit and compare it to the correct voltage on the schematic. When you find a voltage that is much higher or lower than it should be, you have to figure out what could cause it. If you know Ohm's law for voltage, current and resistance, it's not too hard to decide what's causing the undesired voltage change.

When comparing the measured voltages with those given in the instruction book, don't be lead astray by the fact that the instruction book values are "nominal" values – the actual, measured voltages may be 10 per cent higher or lower. This isn't usually a problem in solid-state circuits because the measured voltages should be within 1 or 2 volts of that specified, but in vacuum-tube equipment the measured voltages may be as much as 30 or 40 volts off and still be within the "nominal" range. In a transmitter stage with a "nominal" 800-volt plate supply the actual circuit voltages could fall in the range from 700 to 900 volts and still be okay. The clue here is the actual dc supply voltage, so the first thing to check is the dc supply voltage at the output of the last power supply filter. If it's 10 per cent higher than that noted on the schematic, you can expect other unregulated voltages in the set to be 10 per cent higher.

voltage dividers

We'll discuss series and parallel resistance circuits in more detail next month when we get into troubleshooting by resistance measurement, but in the meantime let's look at a typical series circuit and see what happens when one of the resistors in the string changes value for some reason. Consider the simple series cricuit in **fig. 1**. Since the resistors are connected in series, the same current flows through them all and the resistors divide the voltage in direct proportion to their resistance values. In the circuit of fig. 1 the resistance ratios are 8:4:3:1. The R1-R2 voltage is 24 volts below the supply voltage, the R2-R3 junction is 12 volts below R1-R2, the R3-R4 junction is 9 volts below R2-R3, and 3 volts are developed across R4. The 8:4:3:1 ratio is maintained. (Although circuit voltages are measured in reference to ground unless otherwise specified, you can directly measure the voltage drop across a resistor by placing the voltmeter probes on each lead of the resistor. Be sure the negative voltmeter lead is placed at the lower voltage end of the resistor.)

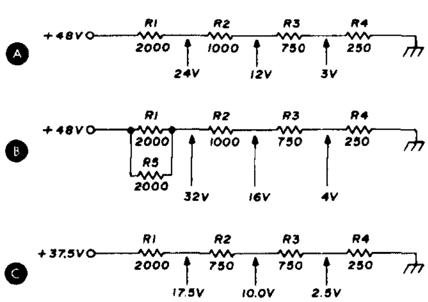


fig. 1. Typical series voltage dividers, showing how the resistance and voltage- dividing ratios are maintained even when a resistor or supply voltage changes value.

Now consider what happens when you change one of the resistance values as in **fig. 1B**. A 2000 ohm resistor has been added in parallel with R1, lowering its effective resistance to 1000 ohms. This changes the resistance ratio of the divider chain to 4:4:3:1, and the voltage

divides accordingly. R1 (with R5) drops 16 volts, R2 drops 16 volts, R3 drops 12 volts and R4 drops 4 volts, maintaining the 4:4:3:1 voltage ratio.

If you forget R1 (and R5) for a moment, note that the ratios of R2, R3 and R4 are 4:3:1 for both circuits. This is important because it illustrates the fact that if one resistor in a series voltage divider changes value, the ratio of the other resistors in the divider remains the same. As a further example of this consider **fig. 1C** where the value of R2 has been reduced to 750 ohms and the supply voltage lowered to 37.5 volts. The voltage and resistance ratios are 8:3:3:1 with the ratios between R1, R3 and R4 the same as in **fig. 1A**.

As an example of voltage troubleshooting, consider the simple voltage divider circuit of **fig. 2**. This is the type of circuit that might be used to provide different operating voltages to various transistor circuits in a set. The bypass capacitors provide necessary circuit decoupling. The circuit of **fig. 2A** shows the normal dc voltages (usually called operating voltages) while **fig. 2B** shows the voltages which you might measure in the circuit when you start troubleshooting.

In **fig. 2A** the +12 volts appears at the junction of R2-R3 because of the 6 volt drop across R1 and R2. A further voltage drop across R3 causes +6 volts at the junction R3-R4. When analyzing the incorrect voltages in **fig. 2B** note that two of the voltages have changed. Since the voltages have changed, it follows that the resistance ratios have changed.

The first step in troubleshooting this circuit, therefore, is to determine what the new resistance ratios are. If you consider only the +4 volts at the R2-R3 junction,

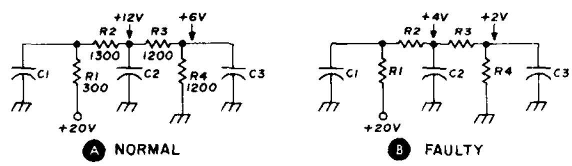


fig. 2. Voltage divider which might be used in solid-state electronic equipment to provide operating voltages to different stages. The incorrect voltages in (B) are easily analyzed with voltage ratios to determine the bad component.

there are two possibilities: R1 and R2 have higher resistance than normal, resulting in a larger voltage drop, or R3 and R4 have lower resistance with a lower than normal voltage developed across them. Which is it? The clue lies in the fact that the ratio between the voltages at R2-R3 and R3-R4 is the same in both circuits, 12:6 = 4.2, or 2:1. Therefore the trouble is more likely in either R1 or R2; one of them has probably increased in value.

You might be inclined at this point to disconnect the two resistors from the circuit and measure them with an ohmmeter. However, further voltage measurements will indicate that one retains its ratio to R3-R4 while the other does not. Even though the voltage isn't shown for the R1-R2 junction, you can quickly calculate it with Ohm's law. Since 6 volts appears across R 4, a 1200 ohm resistor, the current through the circuit is 5 mA. Therefore, the voltage at the R1-R2 junction should be 1.5 volt, a ratio of 1:8 when compared to the voltage at R2-R3.

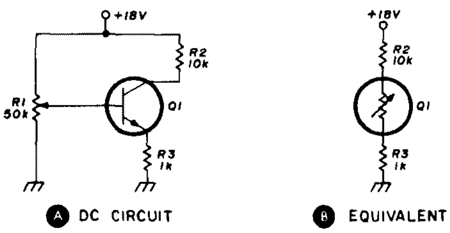


fig. 3. Basic transistor circuit and its equivalent circuit when the transistor is considered to be a variable resistance. Equivalent is complicated because resistance is a function of the base bias which is set by R1.

If the voltage at R1-R2 is 0.5 volt in the faulty circuit, it has the correct ratio to the 4 volts at R2-R3 and resistor R1 is the culprit. On the other hand, if you measure 1.5 volt at R1-R2 in the faulty circuit, R2 has increased in value and should be replaced.

The same type of reasoning is the basis for analyzing all dc voltages in series circuits. First look at the ratio of resistances, compare the voltage ratios, and then figure out what's causing the problem.

transistor and tube circuits

Thinking in terms of resistance and voltage is simple so long as the resistances are simple, and with a little experience you'll be able to estimate voltage ratios close enough to give you a clue to which resistance has changed. Transistor and IC (and vacuum tube) circuits, however, are different because the dc operation of the device changes as you alter the bias and/or supply voltages. In the circuit of fig. 3A, for example, the current through the transistor (and the voltage drop from collector to emitter) is determined by the base bias which is set by R1. Increasing the base bias (base-to-emitter voltage) increases the base current which is multiplied by the current gain of the transistor. This increased emitter current increases the voltage drop across the emitter resistor which affects bias, which affects base current, which affects emitter current, and so on.

The same sort or thing occurs in the vacuum-tube circuit in **fig. 4**. Here the grid bias is picked off the

cathode resistor R1. Since the current through the tube is a function of grid bias, any changes in plate current are reflected throughout the series circuit, which affects bias, which affects plate current, etc.

Although the transistor and vacuum tube can be rather loosely represented by an equivalent variable resistance, the interdependency of bias and emitter or plate current make it difficult to treat active devices as simple ratio dividers. There are some circuit difficulties that can be tracked down with voltage ratios, but you have to be very careful to distinguish between cause and effect. In many cases it is practically impossible to separate the two without resorting to another troubleshooting technique. But don't feel too badly if you get caught in this trap — more than one technician has chased his tail around a circuit only to discover that what he thought was the cause was really the effect, and vice versa.

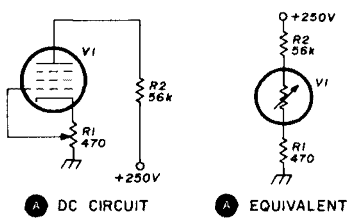


fig. 4. Basic vacuum-tube circuit and its basic series resistance equivalent. Plate resistance is a function of grid bias which is set by the cathode resistor R1. This complicates the analysis as discussed in the text.

Consider the transistor amplifier circuit shown in **fig. 5A** and its resistive equivalent in **fig. 5B**. Voltages are shown for each point in the circuit so you can calculate the voltage and resistance ratios. There is a 9 volt drop across R2, a 6 volt drop across Q1 and a 3 volt drop across R1 so the ratio is 3:2:1. Now assume that something goes wrong with the circuit and you measure the voltages shown in **fig. 5C**. With 4.2 volts across R1, 1.2 volts across Q1 and 12.6 volts across R2 you have a voltage ratio of 3:0.29:1. Since the ratios of R1 and R2 remain the same, it's a good guess that they're okay, but the resistance of Q1 has changed, upsetting the voltages in the circuit. However this doesn't necessarily mean the transistor is bad. It's very likely that something in the circuitry at the base of the transistor (not shown) is causing the problem.

To get an idea of how complex these relationships can get, look at the transistor audio amplifier shown in **fig. 6A**. Shown in **fig. 6B** is the equivalent diagram of the collector circuit ($Q1_C$ is the collector-emitter junction); **fig. 6C** shows the equivalent circuit for the base circuit and includes the bias network (R1 and R2) and the path through the base-emitter junction ($Q1_B$). Note that the emitter resistor R4 is in this path, too, so a current change in either circuit affects the voltage drop across that resistor.

Fig. 6D shows the combined dc paths through the transistor. Figuring out the voltage ratios in this circuit would be difficult even if the resistances were simple, whole numbers, which they aren't, but the circuit is complicated by the fact that the value of $Q1_C$ is controlled by $Q1_B$. However, as will be seen later, there are some rules of thumb that remove some of the apparent complexity and allow you to successfully use voltage ratios to troubleshoot circuits of this type.

We will study transistor circuits in greater detail in a future column, but there are several important facts about transistor circuits that are particularly helpful in understanding the operating voltages of the stage. In a transistor amplifier, for example, the base-emitter junction is always forward biased and the base-collector junction is always reverse biased.* That is, the base terminal is always at a higher dc potential with respect to the emitter, and the collector is always at a higher potential than the base. In npn transistors the collector is positive with respect to the emitter, and in pnp transistor circuits the collector is negative with respect to the emitter. Furthermore, there's an approximately 0.7 volt voltage drop from base to emitter in silicon transistors, and about 0.2 volt base-emitter voltage drop for germanium transistors. In **fig. 6A**, for example, there's a 0.7 volt difference between the base and emitter terminals so Q1 is a silicon transistor.

It's also important when analyzing transistor stages to remember that the base current is a small fraction of the collector or emitter current, typically 1 per cent or less (for a collector current of 25 mA, typical base current is about 250 μA). Therefore, the fact that R4 in **fig. 6D** is common to both the base and collector equivalent circuits is of little consequence because base current contributes only about 11 mV to the voltage drop across R4.

Since the base-emitter voltage remains relatively constant throughout the operating range of the transistor, this can complicate troubleshooting because if the emitter voltage increases for some reason, the base voltage will follow right along behind it. On the other hand, if the base bias voltage increases, this increases the base current slightly, increases the emitter current greatly, and increases the voltage measured at the emitter terminal. The measured voltages may be the same in both cases but the causes are different.

*Forward biased for class-A stages. Class-AB stages are slightly forward biased while class-C stages are operated at zero bias.

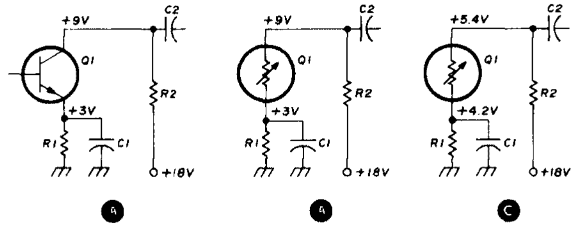


fig. 5. Simple transistor circuit (A), the dc equivalent of the collector circuit (B), and incorrect voltages which can be analyzed using voltage ratios. Troubleshooting cannot be completed however, without considering the base circuit as shown in fig. 6.

The circuits of **fig. 7** are the same as those of **fig. 6A** except that the operating voltages have changed, indicating trouble. Note in both cases that the emitter voltage has increased. In **fig. 7A** the emitter voltage has increased to 1.6 volt while the collector voltage has dropped to 5.4 volts. In **fig. 6A** the ratio of the voltage drops across R3 and R4 is approximately 8:1. In **fig. 7A** the R3 and R4 voltage drops are 12.6 and 1.6 volts respectively, a ratio of about 8:1. Therefore, the difficulty is the base bias network (the value of R1 has probably decreased, increasing the current flow through R1-R2 and increasing the base bias voltage).

In **fig. 7B** the emitter voltage has also increased, but note that the ratio of the voltage drop across R3 (15.6 volts) to that across R4 (1.3 volts) is now 12:1. Further checking will reveal that the value of the emitter resistor has decreased to about 100 ohms, nearly doubling collector and emitter current.

The i-f amplifier in **fig. 8** is typical of the type you might find in a modern communications receiver. Assume you have tracked a receiver problem to this stage and measure the transistor voltages shown in **fig. 8B**. The collector voltage is very low, indicating either higher than normal *current* through the transistor or that R4 has increased in value. The base voltage is a little low, but it has changed little with respect to the emitter voltage. If you study the circuit you quickly decide that the bypass capacitor C3 has shorted. With 1 volt of forward bias the transistor conducts heavily, dramatically lowering collector voltage.

There are countless transistor and IC supply circuits which you can analyze in this same way. First, pick out the voltage that is the most wrong and find out what caused it. Concentrate on one supply path at a time and try to ignore the effects of other circuits. When you decide what happened in one circuit, then decide whether another circuit could possibly be causing the incorrect voltages in the circuit. The component that is common to all symptoms is usually the culprit. Each symptom leads to another, and ultimately to the defective component.

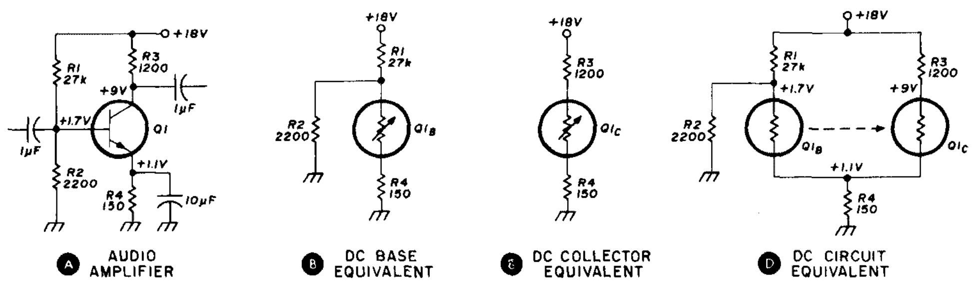


fig. 6. Transistor audio amplifier (A), dc equivalent of the base circuit (B), dc equivalent of the collector circuit (C), and the complete equivalent including collector current dependence upon base current (D).

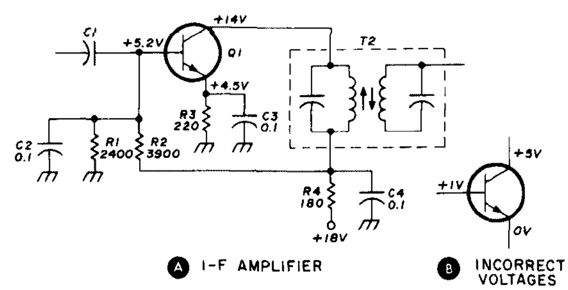


fig. 8. Transistor i-f amplifier with correct operating voltages is shown in (A). Incorrect operating voltages in (B) are analyzed in text.

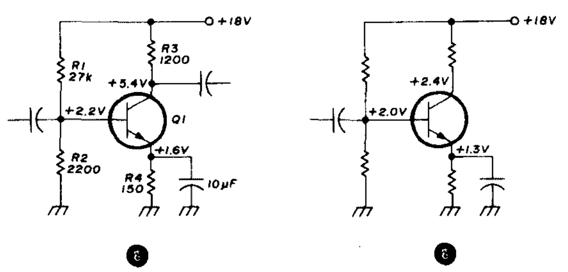
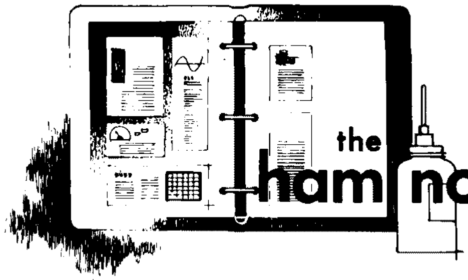


fig. 7. Transistor audio amplifier circuit of fig. 5 with incorrect operating voltages. Although the emitter voltage has increased in both of these circuits, the cause is different in each case as discussed in the text.

Collins paint

Although not directly in the area of troubleshooting, maintaining the appearance of your amateur equipment is also important. Not generally known is the fact that Collins Radio stocks spray cans of paint for both the S-line and the older 75A4/KWS line. The S-line color scheme is actually in three different hues: 180 Gray for the cabinet (Collins part number 097-6161-000), 250 Gray for the panel (Collins part number 097-6162-000) and 126 Medium Gray for the ring (Collins part number 097-6163-000). The spray paint for the 75A4/KWS line is St. James Gray (Collins part number 097-6164-000). Spray cans may be ordered through your local Collins dealer.

ham radio



the ham notebook

high power calibration for the Heath HM2102 vhf wattmeter

The great crowds of two-meter fm enthusiasts will no doubt provide the Heath Company with a continuing market for their HM2102 vhf watt meter. This dandy piece of equipment provides two switch-selectable power ranges of 25 and 250 watts full scale, in the 50 to 160 MHz range, and also includes a built-in swr bridge.

In checking out my wattmeter I found that everything worked fine with the exception of a noticeable discrepancy in the accuracy of the high power (250 watt) range. Checking further, I found that the problem was due to the fact that R8, a 68k resistor which is used as a meter multiplier in the 250 watt range, was out of tolerance. The problem could have been cured by replacing R8 with a new resistor, but it occurred to me that even greater accuracy could be obtained by replacing R8 with a variable 100k resistor to allow separate calibration of the high power range.

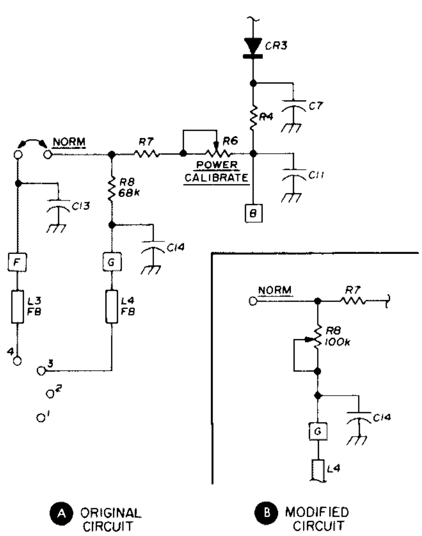


fig. 1. The high-power range of the Heath HM2102 vhf wattmeter can be calibrated by replacing R8 with a 100k potentiometer such as the Allen-Bradley ZV1041 or Bournes 3389P. Calibration procedure is discussed in the text.

This modification is easily done and works out very nicely. The small trimpot may be supported by using short pieces of solid hookup wire inserted in the PC board holes formerly occupied by R8. The adjusting slot should face upward in the same direction as potentiometer R6. A hole in the cover plate allows access to R8 when the cover is in place. This access hole can be labeled high for high power adjust, and the hole already in the plate for R6 can be labeled *low*.

The original calibration procedure calls for adjusting R6, a 50k pot, in the low power position, which also affects the high power calibration. With the addition of a 100k pot for resistor R8, the new calibration procedure is as follows:

1. Using a known power source in the 10 to 25 watt range, and/or comparing with another wattmeter of known accuracy, adjust R6 for the correct meter indication in the 25 watt range.

2. Switch to the 250 watt range, and again using a transmitter with known output, and/or a comparison wattmeter, adjust R8 for correct meter indication.

Both ranges are now individually calibrated. Any further adjustment of R6 will require readjustment of R8.

Robert H. Johnson, W9TKR

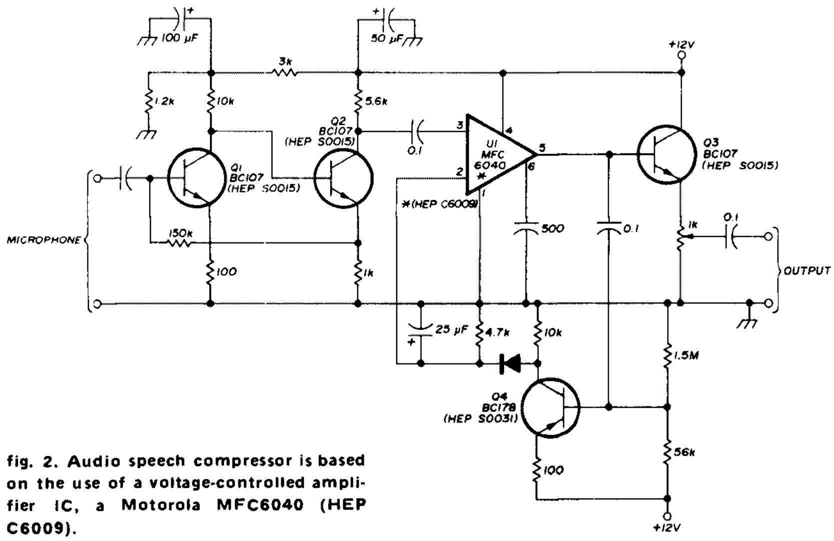


fig. 2. Audio speech compressor is based on the use of a voltage-controlled amplifier IC, a Motorola MFC6040 (HEP C6009).

speech compressor

I wanted to improve the efficiency of my homebrew ssb transmitter, so I decided to build an audio speech compressor. The circuit in fig. 2 uses a Motorola MFC6040 voltage-controlled amplifier IC which has 13 dB gain and 90 dB (maximum) gain reduction. Maximum input is specified at 5 mV rms. In this circuit transistors Q1 and Q2 are a microphone preamplifier. Transistor Q4, which is connected to the output through a 0.1 µF capacitor, is the agc

detector/amplifier for U1, the voltage-controlled gain stage. Q3 is the output buffer.

At my station I use a 500-ohm dynamic microphone with this circuit and the output remains constant at 1.8 volts rms. The performance of the unit could be further improved by adding a 300-3000 Hz filter at the output.

L. Novotny

goral oscillator notes

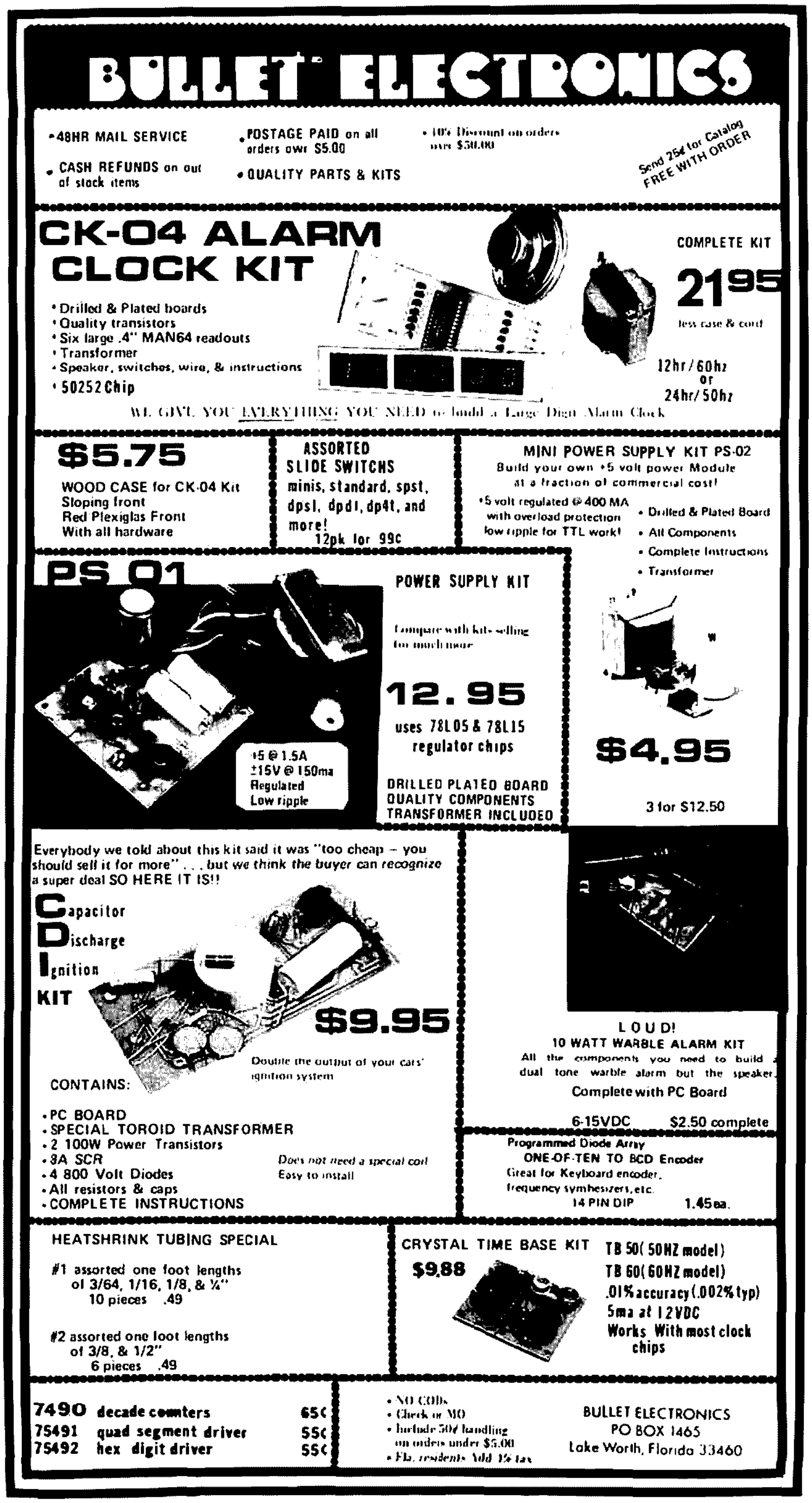
The Goral crystal oscillator circuit described by Don Stoner in *ham radio** appears to be excellent in many respects. I have found, however, that the proper value of C2 in fig. 4 of the original article is a critical function of the capacitance for which the crystal is calibrated. Crystals for the GE Progress Line, for example, are ground to operate into a 10-pF load and will not oscillate on their proper frequency using 20 pF as the value of C2. Data on two different crystals for a GE Progress Line receiver are shown in fig. 1. A value of 12 pF for C2 is more suitable as it allows the crystal to be netted using an 8-pF trimmer capacitor at C1. The data also illustrate the wide frequency range over which the oscillator will operate when different values of C1 and C2 are used.

Robert E. Cowan, K5QIN

nicad battery care

Most pocket computers are powered by rechargeable nicad batteries. These are good batteries, but they must be treated with care. If you run the batteries, or even one cell, much below 0.7 volt, there seems to be the danger of the weakest cell reversing its polarity and chemically burning itself out. If one cell does go dead, it is suggested you replace the

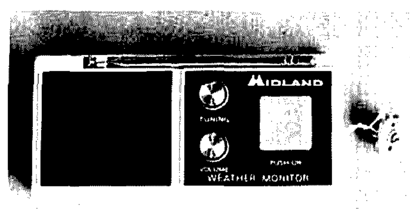
*Donald L. Stoner, W6TNS, "High-Stability Crystal Oscillator," *ham radio*, October, 1974, page 36.



whole string in series, or you may have further problems with cells burning out. If you can get a rundown battery to recharge a little, you may be able to cycle the battery back to health by recurrent discharging and recharging. It appears these cells may also remember how you treat them. Treat them ruggedly, and they will be rugged; treat them kindly, and they wilt away.

monitor receiver modification

I would like to elaborate on the W3WTO article in the January, 1975, issue of *ham radio*.* In this unit the local oscillator is 10.7 MHz *below* the 162 MHz received frequency, at about 151 MHz. By slightly spreading the turns of the oscillator coil, it can be moved to the range of 156 to 159 MHz, 10.7 MHz *above* the two-meter band. The present tuning arrangement covers about 3 MHz, or 145 to 148 MHz if the coil is carefully adjusted.



Weather Monitor tunes from 145 to 148 MHz.

Since the rf coils were previously peaked at 162 MHz, sensitivity on two meters can be substantially improved by *replacing* the rf coils using the same size wire and coil diameter. L1 and L2 should have one additional turn and L3 should have two additional turns. Carefully adjust the length of the new coils for optimum sensitivity.

While the unit works well with its self-contained antenna, I added a phono jack for convenient connection to an external antenna.

It really gives quite good performance for a tunable, $15 two-meter receiver.

Lowell White, W2CNQ

*Kent Mitchell, W3WTO, "Return Weather Monitor Receiver for Two-Meter FM," *ham radio*, January, 1975, page 56.

NEW products

general coverage receiver

The SSR-1 receiver is a new addition to R. L. Drake Company's family of communications equipment. Several design features make it a good candidate for portable work, general-purpose shortwave listening, emergency use, or as a standby receiver. The SSR-1 is frequency synthesized and covers 500 kHz to 30 MHz, providing reception in the a-m, CW, and ssb modes, with selectable upper or lower sidebands.

The SSR-1 is completely self contained including built-in speaker, removable telescoping antenna, 117/234 Vac 50 to 60 Hz power supply, and provision for eight D-cell batteries. With batteries installed, the SSR-1 switches automatically to battery operation if ac power fails. To conserve battery power, the SSR-1 features a front-panel push-button switch that must be depressed to illuminate dial lights.

More information may be obtained by writing to the R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342, or use *check-off* on page 110.

up/down counters

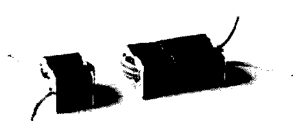
ESE is now producing the ES-301 and ES-302 digital up/down counters. Both are four-digit, 100-minute timers featuring four gas-discharge displays for display up to 99:59. Six separate controls count up, count down, stop, minutes advance, seconds advance and reset. The controls are momentary push-button switches. When the stop control is pressed, the display is automatically held at the precise second. Both the ES-301 and the ES-302 may be preset to a desired number for a specific count, and timing can be activated from that point, up or down. Desired numbers on the ES-301 can be preset by advancing the minutes and seconds simultaneously or independently. Lever wheel type switches instantly preset the number on the ES-302.

Depressing the reset button on both units returns the numbers to 00:00 from which they will continue counting up or down, unless the stop button is pushed. Both units may be equipped with an option that returns the number to the preset digits when the reset is activated. Counting direction (up or down) on both units can be reversed or reset to 00:00 without stopping the count.

Both the ES-301 and ES-302 come in an etched aluminum case with simulated walnut sides and top. Power for both is 7 watts maximum, 117 Vac at 60 Hz. The ES-301 and ES-302 are efficiently designed for constant, daily use, utilizing solid state reliability, silence, easy operation, high accuracy, long life, low initial cost and operation.

For detailed catalog sheets contact ESE, 505½ Centinela Avenue, Inglewood, California 90302 or use *check-off* on page 110.

wideband rf transformers



Communications Power is offering a complete line of wideband rf transformers designed specifically for impedance matching in high-power solid-state amplifiers. The transformers cover 1.8 to 30 MHz and are rated at 150 watts. Extremely rugged construction assures reliability in any environment. Applications include marine as well as military and amateur radio communications equipment.

High-volume manufacture means lowest prices. The following example is representative of CPI wideband rf transformers available for immediate delivery. These units all have turns ratios of 1:3, 4, 5, or 6 and cover 1.8 through 30 MHz:

series	power rating (watts)	price (1 - 4 pieces)
RF1000	150	$5.00
RF800	100	4.00
RF600	50	3.50
RF400	25	2.50

Information on other types is available on request. Write Communications Power, Incorporated, 2407 Charleston Road, Mountain View, California 94043, or use *check-off* on page 110.

fm signal generator

The Edison Electronics division of McGraw-Edison Company has developed a solid-state fm signal generator that covers all the mobile communications frequency bands allocated by the FCC. Four models are offered, designed to your specific carrier-frequency needs. Each model has six frequency bands. The model 800A covers 25 to 960 MHz; model 801A, 25 to 470 MHz; model 802A, 25 to 175 MHz; and the model 803A, 25 to 520 MHz. Any desired frequency can be quickly obtained by first selecting one of the six frequency bands, then tuning the coarse tuning control until the desired frequency appears on the hand-calibrated tuning dial. Finally, narrowband adjustments may be made with either an electronic fine tuning control or incremental frequency controls.

Output voltages are provided with accuracy traceable to NBS. Output is continuously variable between 0.1 microvolt and 0.1 volt. A temperature-compensated bolometer circuit maintains output voltages automatically. Accurate receiver sensitivity measurements can be made to 0.1 microvolt.

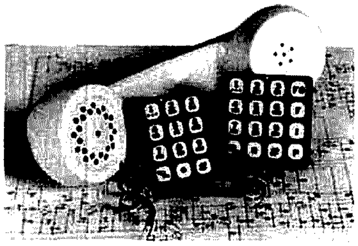
The Measurements Model 800A series fm signal generators feature internal modulators that provide fm at 1000-Hz

More Details? CHECK—OFF Page 110

sine waves or 20-Hz sawtooth waves. External modulation between dc and 30 kHz may be applied through binding posts on the panel. Sync output and sync phase are available for external modulation up to ±32 kHz peak deviation so that dual-trace sweep alignment may be used.

All four models are available at $992,00 FOB Manchester, New Hampshire. For a brochure providing more technical details write Edison Electronics, Grenier Field, Manchester, New Hampshire 03103 or use *check-off* on page 110.

tone encoding keyboards



Four new tone encoding keyboards have been introduced by Electrografix for vhf/uhf installations where access is required to amateur autopatch repeaters. Designated TEK-125, -165, -225, and -265, the series incorporates the cmos IC developed by Motorola: the MC14410 digital tone encoder.

The pads provide a compact, accurate, low-power, digital tone encoding system with a full 2-of-7 or 2-of-8 encoding format from a basic 1-MHz crystal oscillator. A unique key pad switch complements the anti-falsing lockout feature of the Motorola IC.

The two smaller pads, TEK-125 12 button and TEK-165 16 button, are for use with hand-held transceivers or other small units. TEK-225 and TEK-265 are much larger and are intended for installation in remote-control panels, repeater sites, or on vehicle dash panels. All units are 0.40 inch (10.2mm) thick. External dimensions are: TEK-125, 1.58x2.08 inches (40x52.8mm); TEK-165, 2.08x2.08 inches (52.8x52.8mm); TEK-225, 2.05x2.70 inches (52x68.6mm); TEK-265, 2.70x2.70 inches (68.6x68.6mm).

Also featured are a glow-in-the dark keyboard face and a LED in the bezel

face to indicate when a tone has been generated. The LED also functions as a battery-level indicator. The keyboard face is nonradioactive, and when exposed to normal sunlight or other similar light, it will glow up to eight hours. The phosphor green color is highly visible and legible under all lighting conditions, from bright sunlight to total darkness. Other colors are available in quantity purchases.

The TEK series tone-encoding keyboards have gold-plated circuit boards, single-unit molded ABS plastic bezels and cases, and an externally adjustable level control. Combined operating current for the tone generator and LED indicator is less than 13 mA. When operating as a battery-level indicator, the LED current drain is less than 8 mA at the rated input of 6 to 16 Vdc. Output level is a 0 to 600 mV p-p composite waveform, which will modulate any transmitter. The TEK-165 weighs less than 0.9 ounce (27g).

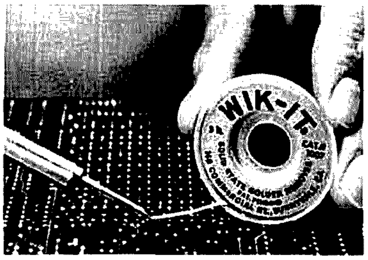
The TEK keyboards are complete and ready to go. Only three electrical connections are needed at the transmitter: audio, B+, and ground. Mechanical installation is simple – either by self-tapping screws or adhesive strips provided with each unit.

The single-unit list price for TEK-125 is $57.50; for TEK-165 it is $65.00. The TEK-225 and 265 prices are available on request. For additional information, write Electrografix, Inc., P.O. Box 869, Chino, California 91710, or use *check-off* on page 110.

semiconductor replacement guide

This book is designed to fill a gap in the information available to amateurs and electronics technicians. It provides general-purpose replacements for manufacturers' semiconductor parts numbers. Over 15,000 semiconductors used in entertainment-type electronic equipment are cross-referenced to the universal replacements produced by General Electric, International Rectifier, Mallory, Motorola, RCA, Sprague and Sylvania. Included are bipolar and field-effect transistors, diodes, rectifiers, and integrated circuits. 256 pages, softbound, $3.95 from Ham Radio Books, Greenville, New Hampshire 03048.

solder remover



You can de-solder any soldered joint merely by placing your iron atop a special braid, called *Wik-It*, which in turn is laid on the soldered joint. In a second or two the solder simply disappears into the special braid. The soldered joint is now clean and free of solder, ready for the next operation.

Wik-It is a patented solder remover developed by the Wik-It Electronics Corporation. Because of special chemical treatment of the braid, the solder to be removed is drawn into the braid through capillary action. When the solder has been removed from the joint, the user just snips off about a half-inch of *Wik-It* which is then ready to be used again.

Because the solder removal occurs so quickly, little heat is transferred to the surrounding material, whether it be mounting board or wire insulation. As a result, *Wik-It* eliminates delamination. lifted pads, and measling which are all too often seen with de-soldering attempts. Flux contamination and component damage are also eliminated by the quickness of the *Wik-It* method.

Wik-It has been thoroughly tested for over two years in manufacturing companies and is approved for use in military and aerospace work since it meets specification MIL-F-14256C Type W and Type A.

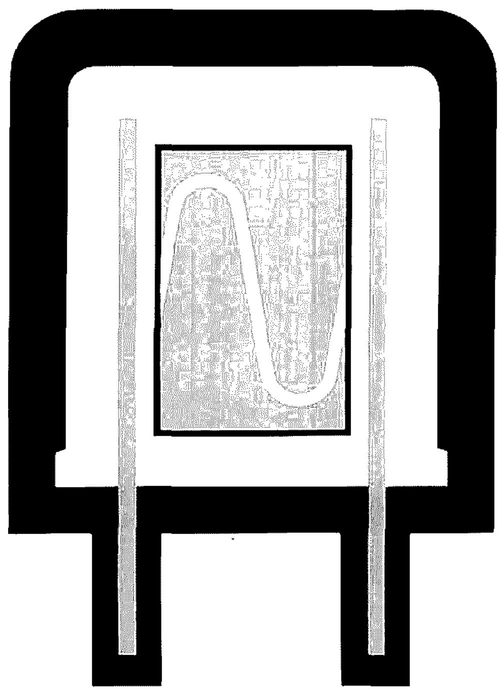
Wik-It is available in different sizes to suit small transistor and IC applications as well as larger tube and heavier-size wire work. *Wik-It* comes in various lengths from 5 to 100 feet. The price of a 5-foot roll of *Wik-It* is from $1.59 to $1.79, depending on width, and is available in electronics supply stores or from the manufacturer. Free samples are available upon request.

For more information, contact *Wik-It* Electronics Corporation, 140 Commercial Street, Sunnyvale, California 94086, or use *check-off* on page 110.

one dollar

ham radio magazine

MARCH 1976



survey of
crystal oscillators

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offices
Greenville, New Hampshire 03048
Telephone: 603-878-1441

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Ham Radio Europe
Box 444
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United Kingdom
Ham Radio UK
Post Office Box 64, Harrow
Middlesex HA3 6HS, England

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Republic of South Africa

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contents

a second look

by Jim Fisk

Recently there has been increased interest in the work of Nikola Tesla, one of the most remarkable inventors of all time, and a man who has rightfully been called the "father of radio" by some of his contemporaries. Born in Yugoslavia 120 years ago, Tesla arrived in America in 1884 with only four cents in this pocket and a letter of introduction to Thomas Edison. Edison immediately put the new immigrant to work in his New York laboratories, but the methods and temperament of the two men were fundamentally different, and the arrangement lasted less than a year. Tesla was a lone wolf, secretive, tried nothing which he had not previously thought out to the last detail, and believed that alternating current was the electrical power of the future. Edison, on the other hand, preferred to work by trial and error and was totally committed to the direct current which powered the incandescent lamps which he invented and his company lighted.

After leaving Edison's employ Tesla worked as a common laborer – the only job he could get – but his fortunes changed in 1887 when he persuaded some financiers to underwrite his own Manhattan laboratory. His multiphase ac machinery, the designs for which he had been carrying around in his head for five years, was built, tested, and patented before the end of the year. George Westinghouse, Edison's competitor, reportedly paid Tesla a million dollars for the patents, but after Tesla's death in 1943 a personal friend said that Tesla actually got only $200,000, and two-thirds of that went to his financial backers. Considering that practically all modern alternating-current machinery – generators, motors, converters, regulators and transformers – are based on early Tesla patents, the price was a pittance.

Tesla's best known invention is probably the Tesla coil, the high-frequency, air-core oscillation transformer which is still used in laboratories to demonstrate high-voltage phenomenon. The largest of these, which Tesla built in Colorado in 1899, was used to light 200 50-watt lamps at a distance of 26 miles without any connecting wires. Convinced of the possibility of wireless power transmission, Tesla began construction of his first demonstration plant on Long Island in 1904. Shortly after the massive, 187-foot octagonal tower was completed, however, he ran out of money and was forced to abandon the project. Since Tesla never revealed how large an area he intended to cover and he kept few notes, little is known about the installation. However, this and other of his works are receiving renewed interest from modern energy researchers.

Not so well known are Tesla's many inventions in the fields of lighting, turbine engineering, automation, high-frequency alternators, Xray apparatus, induction heating, and radio communications. One of his neglected inventions, still unused, was a carbon-button lamp which gave twenty times more light, for the same amount of current, as Edison's incandescent filament lamp. He also discribed a system for detecting ships and other distant objects by aiming a powerful beam of short-wave impluses at them and picking up reflections on a fluorescent screen – a clear prophesy of the radar of the future.

Tesla first started working with high-frequency energy in the late 1880s, and at a lecture in 1892 he demonstrated wireless communications circuits which contained all the basic elements of those adopted several years later by Marconi and others. These circuits remained the same for all radio communications until the introduction of the transmitting electron tube nearly three decades later.

Believing that point-to-point communications was an obvious application of Hertz's own 1887 experiments, Tesla made no attempt to patent his very early wireless apparatus. He was primarily interested in developing a new and efficient method of wireless power transmission, and although he was never completely successful, he developed and patented some of the most advanced wireless apparatus of the day, apparatus which was eagerly adapted by others for more profitable purposes.

In 1915 Tesla sued Marconi for infringement of his patent rights. In the long and drawn out court battle that followed Tesla became more and more of a recluse, poverty stricken, moving from one hotel to another when he wasn't able to pay his bills. Ironically, shortly after his death the Supreme Court finally declared that Marconi's four-circuit wireless patent (his most important) was invalid – it was predated by the work of Tesla, John Stone, and Oliver Lodge. Both Stone and Lodge acknowledged that their inventions had been inspired by the *earlier* lectures of Tesla. Perhaps on this 120th anniversary of his birth Tesla will begin to receive some of the recognition he deserves – without his genius we would all be the poorer.

Jim Fisk, W1DTY
editor-in-chief

presstop

A NEW WARC TASK FORCE for "Research and Support" was announced by Chairman John Johnston as the first item of business at the January 20th WARC '79 meeting in Washington. This Task Force will be headed by Dick Baldwin and will supply (through ARRL) the basic working group support for Amateur Radio WARC preparation he had promised at the December meeting.

The Proposed Frequency Table was then reviewed by the Task Force chairmen with no significant changes from those previously proposed (Presstop, February, 1976 issue). Asking and getting are different things, however. Other U.S. groups are reported to be after 3.9-4.0, 7.2-7.3, 146-148, and 220-225 MHz (see below), and broadcast would like to extend their a-m band upper limit from 1605 to 1805 kHz!

CB'S WARC MEETING on January 13th generated frequency recommendations of one MHz between 26-28; 5 MHz between 216-230 (220-225 MHz preferred), 10 MHz between 470-947 (around 900 MHz preferred), and 200 MHz between 15 and 25 GHz. Present Class-C (72 MHz) and Class-A (462/467 MHz) would remain unchanged.

SIGNING "MOBILE" OR "PORTABLE" with your call will never again be required if a just released Notice of Proposed Rule Making becomes part of the Amateur rules. Docket 20686 also proposes that the requirement for notifying the FCC before engaging in an extended period of portable operation be deleted.

Comment Due Date is February 27, with Reply Comments due on March 8. Though Docket 20686 would make it unnecessary to indicate portable or mobile operation, it would not prohibit such identification — a generally favorable Amateur response is expected, but such a rules relaxation could cause confusion in contests and for award programs.

FCC'S NEED FOR ENFORCEMENT "TEETH" highlighted FCC Chairman Richard Wiley's message to Congress in an appearance before the Senate Commerce Subcommittee January 20. He'd like to be able to double the size of the fines the Commission can levy, extend its authority to cover unlicensed as well as licensed radio operators since under the present rules unlicensed operators — principally CB violators — must be turned over to the Justice Department for prosecution after the FCC has tracked them down.

TRANSMITTER SPURIOUS OUTPUTS such as those previously reported from the Heathkit 2026 and early Multi-2000 are not the only problems facing buyers of vhf/uhf Amateur gear. Because of the potential interference to the many services using the higher frequencies, all receivers that tune above 30 MHz that are sold commercially must be FCC certified that they meet the limits on radiation specified in Part 15 of the rules.

Some Japanese Imports have come in without Part 15 certification and reports are that some uncertified equipment is being sold currently on the Amateur market.

CANADIAN AMATEURS requesting reciprocol privileges no longer have to have a U.S. mailing address — the change, which also applies to CB licenses, is effective immediately.

UNLICENSED AMATEUR-LIKE ACTIVITY by the "HF gang" operating on ssb in the region between channel 25 and 27.5 MHz grows at an ever increasing pace. Latest report is that 10,000 each of HF- and HFA- series "calls" have been assigned to these illegal stations and they are now well into the HFB- numbers! With rumors abundant that Class-D CB expansion to 27.5 MHz is due out shortly, it's not at all unlikely that we'll soon find more of them than licensed Amateurs on the 10-meter band

Amateurs Can Assist FCC in their efforts to cope with this illegal tide by reporting to the FCC any information they pick up listening to the HFers that might help in identifying individual operators — some have been heard recently giving their complete street addresses for QSL purposes, and since their's is an entirely illegal operation, the restrictions concerning secrecy of communications do not apply. Reports can be sent to the Engineer In Charge of your nearest FCC Field Office, but better yet would be one of the four FCC Special Enforcement Facilities with a carbon to your local EIC. Write Supervisor, FCC SEF at P.O. Box 1588, Grand Island, Nebraska 68801; P.O. Box 65, Powder Springs, Georgia 30073; 2914 W. Edinger Avenue, Box 5126, Santa Ana, California 92701; or P.O. Box 36, Laurel, Maryland 20810.

HIRAN DOCKET proposing shared use of 420-450 MHz band by the off-shore oil rig radio location system has finally made it into the home stretch and will be out very soon. Reports are that despite strenuous Amateur and Citizens Division opposition HIRAN will get to share the band on a temporary, non-interferring basis as Docket 20147 had initially proposed.

RCA WILL STOP TUBE PRODUCTION and close its historic Harrison, New Jersey plant July 30th. However, RCA will continue to make some types overseas and buy others from other makers so does plan to continue as a major tube supplier for some time to come.

survey of crystal oscillators

A comprehensive review of many circuits with recommendations to help you choose the crystal oscillator best suited to your design needs

Crystal oscillators are **fundamental** to radio-communications equipment and, today, to many digital devices such as clocks, frequency meters, and some precision measuring instruments. Crystal-oscillator circuits abound, and design choice depends on the application and desired characteristics.

This article presents an overview of solid-state crystal oscillators, many of which have been gleaned from the literature. Unfortunately, many published circuits have unknown or little-understood characteristics, whereas others have inherent pitfalls that can trap the unwary. A major purpose of this article is to ferret out these questionable crystal-oscillator circuits and offer guidelines on what to expect if construction is contemplated.

Fundamental-frequency and overtone-frequency crystal oscillators are reviewed. Applications are presented for low- and high-frequency circuits, 1-MHz time-base clocks, TTL-compatible oscillators using ICs, variable-frequency oscillators (VXOs), oscillator-multipliers, and multichannel (switchable) oscillators. The article is prefaced with a short discussion on quartz-crystal resonant modes, which is necessary to an understanding of how these crystals operate in an oscillatory circuit.

It's sometimes desirable to know something of the various crystal cuts for a particular application; such a discussion is beyond the scope of this article. References 1 to 8 contain suitable material; references 1 to 5 are particularly recommended.

One thing little appreciated about quartz crystals is that the crystal can oscillate in two resonant modes: parallel and series. The two resonant frequencies are separated slightly, typically 2 to 15 kHz. The series-resonant frequency is lower than the parallel-resonant frequency. Oscillator circuits are designed so that the crystal oscillates in one of these modes. Few articles note that a crystal specified and calibrated for use in a parallel-mode circuit may be satisfactorily used in a series-mode circuit if a capacitor equal in value to the circuit's specified parallel-mode load capacitance (usually 20, 30, 50 or 100 pF) is placed in series with the crystal. Unfortunately, the converse isn't true. The series-mode crystal will oscillate above its calibrated frequency in this case, and it's usually impossible to pull it down sufficiently with capacitive loading.

Overtone crystals always operate in the series mode, usually on the 3rd, 5th or 7th overtone, and manufacturers usually calibrate the crystal at the overtone frequency, *not* at the fundamental frequency. Operating a crystal in the parallel mode and multiplying the frequency three or five times doesn't produce the same result as operating the same crystal in the series mode on its third or fifth overtone. When ordering overtone crystals, avoid confusion and specify the frequency you want, not the apparent fundamental frequency!

For crystals used in parallel-resonant-mode oscillator circuits, the load capacitance must be specified, as stated above. Take into account device input capacitances, circuit capacitance across the crystal, and strays. A load capacitance of 30 pF is the most usual specified value. A trimmer is often used in parallel-mode circuits for exact frequency setting and is a desirable feature. When ordering crystals to be used in the series mode, a load capacitance is not specified.

Crystal oscillator circuits can be divided into two modes of operation: fundamental-frequency oscillators and overtone oscillators. Fundamental-frequency crystals cover the frequency range to 20 MHz, although fundamental-frequency crystals above 15 MHz are uncommon and very fragile. Fundamental-frequency crystals are usually specified for parallel-mode operation, but series mode can be requested. Low-frequency crystals, below 500 kHz, are often best operated in the series mode. Overtone crystals cover 15 to 150 MHz.

By Roger Harrison, VK2ZTB, 14 Rosebery Street, Balmain NSW 2041, Australia

Overtone crystals above 100 MHz are usually fragile and expensive but are handy for uhf oscillator-multiplier chains.

The choice of a circuit will depend on your application. Various applications will require sine-wave output, square-wave output or a harmonic-rich output, for example. Other applications may require an oscillator to accept crystals covering a wide frequency range or an oscillator to drive a TTL digital device.

In any crystal oscillator circuit it's desirable to have some means of setting or trimming the frequency and to be able to adjust the feedback, easily and in a noncritical manner, to limit crystal power dissipation. Being able to control the feedback has other advantages, as we shall see later. The most convenient way to adjust feedback is to vary resistors or capacitors. Coil taps and tickler windings don't provide the same ease of fineness of adjustment.

The permissible maximum power dissipation of crystals in the 1 to 20 MHz range, operating in the fundamental mode, is about 200 microwatts and is similar for overtone operation. Most low-frequency crystals, below 1 MHz, have a permissible maximum power dissipation of 100 microwatts and should be operated with a dissipation below 50 microwatts. Operating a crystal above or near this level degrades its stability, and the circuits chosen here avoid this problem by limiting the dissipation.

The equivalent crystal series resistance, which determines its activity, determines crystal power dissipation. The higher the series resistance, the lower the permissible power dissipation. Reference 5 gives a representative list for those interested. Using an 807 tube with 600 volts on the anode in a Pierce crystal oscillator is not recommended -- you'll probably fracture the crystal. The main point to remember is that crystal oscillators are meant to provide a stable frequency source, *not* power output. Frequency stability, both short and long term, and crystal life are compromised when an oscillator is operated at an excessive power level. In all applications it's wise to use a regulated source of supply voltage.

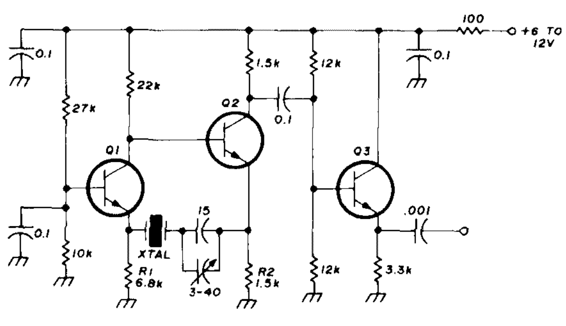


fig. 1. Aperiodic (Butler) oscillator; a series-mode oscillator with sine-wave output. Reducing R2 to about 1k will produce good harmonics to 30 MHz with a 100-kHz crystal. Transistors Q1, Q2 and Q3 are 2N918, 2N2222, 2N3563, 2N3564, 2N3693, 2N3694, 2N5770, AY1119, BC107, BC108, BC109, BC547, BC548, BC549, or SE1001.

aperiodic oscillators

These circuits don't use tuned circuits and can operate over a very wide frequency range. The only component that requires changing for a change in frequency is the crystal. This can be a very useful advantage, whether the circuit is used simply as a "crystal checker" or in some other application. For low-frequency crystals, tuned circuits tend to be bulky, and an aperiodic oscillator has a distinct advantage. However, such oscillators aren't without drawbacks. Some low-frequency crystals, particularly DT and CT cut, are prone to oscillation on unwanted modes.

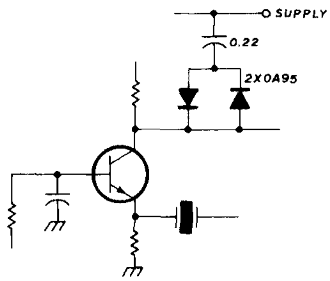


fig. 2. Adding diodes to the circuit of fig. 1 limits oscillator amplitude and hence crystal power dissipation; starting performance is also improved.

When using an aperiodic oscillator, it's wise to check that the output is on the correct frequency and that no mode instability exists. With these crystals, the *esr* (equivalent series resistance) of higher-order oscillation modes is often less than the fundamental-mode *esr*. As a consequence, the crystal oscillates more readily in the undesired mode. One way around this problem in an aperiodic oscillator is to use a transistor with high small-signal gain at low frequencies that decreases rapidly above the desired operating frequency. In extreme cases, an oscillator with a tuned circuit may be necessary.

The circuit in fig. 1 is an aperiodic Butler oscillator. The basic circuit was first published in *VHF Communications*[9] and has since appeared in various versions in other amateur publications. A similar circuit is discussed by M. Lane[6] (see also fig. 3). The output of the circuit in fig. 1 is essentially a sine wave, although the second and third harmonics are not much attenuated. Reducing the Q2 emitter resistor increases the harmonic output. Reducing this resistor to about 1k will produce good harmonics to 30 MHz from a 100 kHz crystal. Being a Butler oscillator, it's a series-mode circuit. Oscillation is noncritical with supply voltage variation, although frequency stability is affected. A regulated supply is recommended for applications where stability is important.

As indicated, a variety of transistors may be used. For crystals above 3 MHz, transistors with a high gain-bandwidth product (f_T at least 100 MHz) are recommended. For crystals in the 50 to 500 kHz range, transistors with high low-frequency gain, such as the 2N3565 or BC549, are recommended. Low-frequency crystals, as mentioned before, have a permissible power dissipation limit of 100 microwatts, and amplitude limiting may be necessary. Low supply voltage, consistent with reliable starting, is one way of achieving this. The addition of diodes to the circuit, as in fig. 2, is probably a better

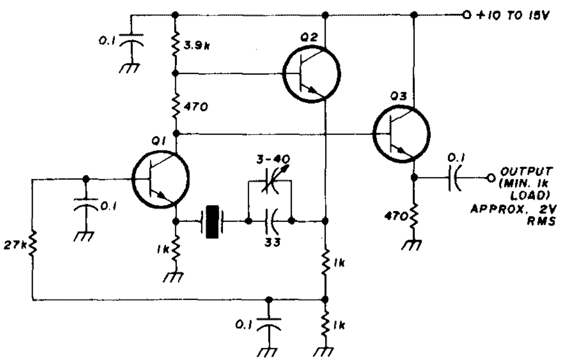


fig. 3. Aperiodic emitter-coupled oscillator (after Lane, reference 6). This circuit is somewhat frequency sensitive to power-supply voltage variations and load changes despite buffered output. Transistors Q1, Q2, and Q3 are 2N3646, 2N3013 or MPS3646.

method, and starting performance is improved. The circuit will oscillate with fundamental-mode crystals to at least 15 MHz with appropriate values of R1 and R2 as well as the appropriate transistors. As shown in the circuit, an emitter follower (or source follower) is recommended.

The aperiodic oscillator in fig. 3 is discussed by Lane,[6] and comments similar to those above apply here. It is also slightly frequency sensitive to power-supply voltage variations. Load changes also affect frequency stability slightly despite the buffered output. A minimum load of 1k is desirable.

The two oscillator circuits just described are series-mode oscillators adapted to accept parallel-mode crystals. When using a known series-mode crystal, the trimming capacitance in series with the crystal should be replaced by a short circuit.

An aperiodic parallel-mode oscillator is shown in fig. 4. It's a wideband dc amplifier, with the crystal providing feedback. A crystal operating in the parallel mode has a 180-degree phase reversal across it; consequently, here it provides positive feedback. The 3 to 30 pF trimmer from Q1 base to ground is for frequency setting. A buffer is recommended. The output is not a pure sine wave and the harmonic level is fairly high.

There are many other aperiodic oscillator circuits, but the three discussed here have proven to have repeatable characteristics. Most IC TTL oscillators are aperiodic, but their application is in digital equipment generally, and they are covered in a later section.

low-frequency oscillators

Crystals in the 50 to 500 kHz range are commonly CT or DT cut, but regardless of the cut, they require special considerations not encountered with the more common AT- or BT-cut crystals used in the high-frequency range. As mentioned previously, their *esr* is usually high and they are prone to oscillating in a higher-order mode, usually at twice the fundamental frequency.

The circuit in fig. 5 has two advantages. It doesn't require a tuned circuit and you have a choice of either sine- or square-wave output. For crystals in the 20 to 150 kHz range, 2N3565, 2N2920 or 2N2979 transistors are recommended to avoid mode-instability problems. Any of the other types listed are satisfactory for crystals in the 150 to 500 kHz range. Frequency stability and mode stability are good, and the circuit makes an excellent oscillator for troublesome FT241 crystals. If the crystal won't start reliably, it probably has a high *esr*. In this case increase R1 to 270 ohms and R2 to 3.3k. For square-wave operation, C1 is 1 μF or a standard value either side. Do not use an electrolytic capacitor. C1 is deleted for sine-wave operation. Harmonic output for sine-wave operation is quite low; the second harmonic is typically −30 dB or more.

A somewhat simpler, parallel-mode, low-frequency oscillator is shown in **fig. 6**. It makes an excellent bfo for 455 kHz. Resistor R_f controls the feedback. If the oscillator won't start, reduce the value of R_f. Increasing the value of R_f reduces harmonic output, a very handy feature. However, at the reduced level of feedback, the oscillator can take up to 20 seconds to reach full output. Harmonic output can be reduced to better than −40 dB.

To produce an output rich in harmonics, bypass the transistor 100-ohm emitter resistor with a 0.1 μF capacitor. Output will rise to about 3 volts rms. Power supply voltage is best kept below 9 volts in this case. For crystals having a specified load capacitance of 30 or 50 pF, remove the 100 pF capacitor in series with the crystal.

Another simple circuit is shown in **fig. 7**. It has a minor disadvantage in that it requires a coil — usually a bulky component. No coil data is given, but proprietary prewound coils of the indicated inductance range are an excellent solution. This circuit allows a simpler arrangement than that of fig. 6, if switched crystals are desirable. The switch contacts should be inserted between C1 and L1 with each crystal having its own coil. Intercontact capacitance across the switch contacts should be low. Adjusting the slug in L1 pulls the crystal frequency.

The circuit of fig. 7 is basically a series-mode circuit. Parallel-mode crystals can be used by making C1 equal to the specified load capacitance as mentioned earlier. Performance is similar to the circuit of **fig. 6**; harmonic output is usually better than −30 dB. CT- and DT-cut crystals are usually used to cover 100 to 500 kHz. These

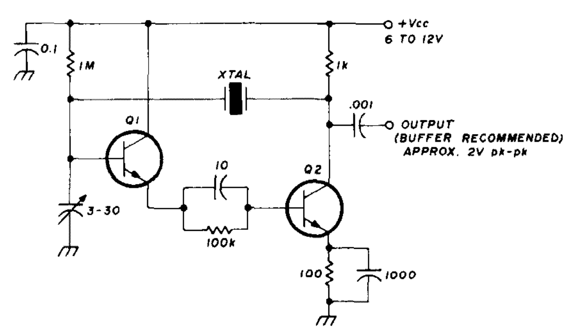


fig. 4. Aperiodic parallel-mode oscillator. The trimmer capacitor sets the frequency. A buffer is recommended to overcome loading problems. Q1 and Q2 are 2N914, 2N918, 2N3565, 2N5770, BC109 or BC549.

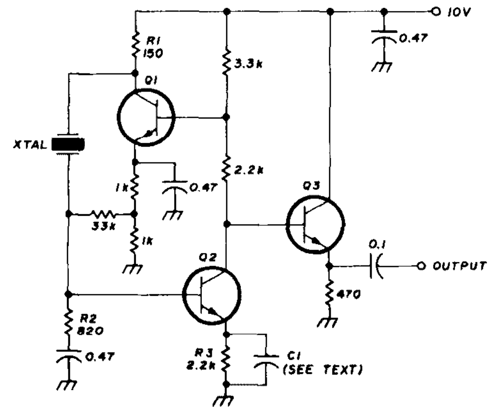


fig. 5. A low-frequency series-mode oscillator (after Lane, reference 6). No tuned circuit is required and either sine- or square-wave output is possible. Circuit is excellent for troublesome FT241 crystals. Output is approximately 1.0 to 1.5 volt rms sine wave or 4-volt square wave (see text).

for 20-150 kHz crystals Q1, Q2, Q3	for 150-500 kHz crystals Q1, Q2, Q3
2N3565	BC107, BC547
2N2920	2N3565
2N2979	2N5770
	2N2222

crystals are prone to oscillating at twice their fundamental frequency; for this reason, the circuit in fig. 7 is recommended where such trouble may be experienced.

For crystals in the 40 to 100 kHz range, the +5-degree X cut is common. The GT cut was once common for this range (up to 250 kHz), and surplus crystals may be of this type. The circuit in fig. 8 is recommended for these crystals or where really cantankerous low-frequency DT-cut crystals are encountered. To trim the frequency of series-mode crystals, tune the slug in L1. For parallel-mode crystals, C2 may be replaced by a capacitor about two-thirds the specified load capacitance in parallel with a 3 to 30 pF trimmer. The trimmer then serves as the frequency adjustment and L1 is peaked for maximum output. The use of a buffer is recommended.

A circuit seen commonly in the literature is that of fig. 9. It works quite well, most of the time, but suffers from a number of disadvantages. Feedback is dependent on the gate-drain capacitance, a highly variable quantity; stability suffers. It is prone to a) frequency pulling with load changes despite good buffering, and b) output-level variations. Performance is not always repeatable. Starting performance varies greatly between fets of the same type and with various crystals.

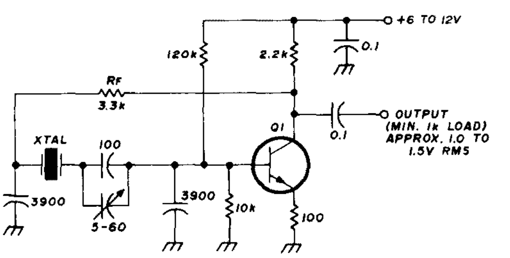
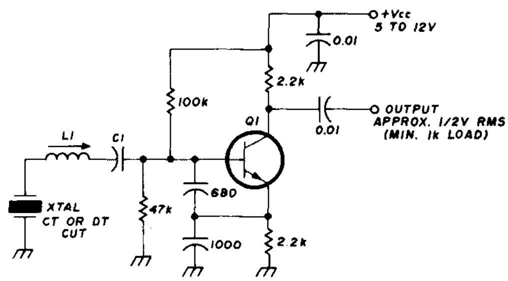


fig. 6. Low-frequency (50 to 500 kHz) parallel-mode oscillator (after Lane, reference 6). Feedback-control resistor R_f may be decreased for hard-to-start crystals; increasing R_f reduces harmonic output. Q1 is a 2N2920, 2N2979, 2N3565, 2N3646, 2N5770, BC107 or BC547.

1 MHz oscillators have many applications, such as frequency markers, clock oscillators and time-base references. The circuit in fig. 10 is recommended where a high-stability 1 MHz source is required. It has a basic stability of 1 part in 10^{-6} or better with supply variations and the usual atmospheric-temperature variations. Using an inexpensive, commercially available crystal oven, it can maintain a stability approaching 1 part in 10^{-8} per day. Harmonic output is low. The output level is about 2 volts peak-to-peak. A TTL-level output is readily obtained by driving a 7413 Schmitt trigger. The circuit originally appeared in *Electronics*.[10]

A simple, general-purpose, parallel-mode, 1-MHz crystal oscillator, fig. 11, uses circuit constants for fundamental crystals in the 800 kHz to 3 MHz range. Output is a sine wave at about 0.5 volt rms into a 1k load.

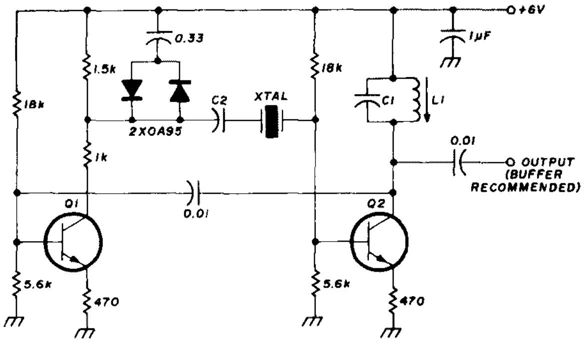


mode	C7	freq (kHz)	L (μH)
series	.01 μF		
parallel	use specified load capacitance (30, 50 or 100 pF)	150-300	800-2000
		300-500	360-1000

fig. 7. Low-frequency (150 to 500 kHz) oscillator (after Foster and Rankin, reference 7). Adjusting slug in L1 pulls crystal frequency. Circuit is basically series mode, but parallel-mode crystals can be used by making C1 equal to the specified load capacitance. Q1 is a 2N3563, 2N3564, 2N3693, BC107, BC547, or SE1010.

Harmonics are better than −35 dB down, although this is dependent on the transistor and the power-supply voltage used. Note that C1 differs for different crystal-load capacitances.

Integrated circuits are coming into increased use in communications equipment. Their versatility and performance present obvious advantages over discrete circuits. The circuit of fig. 12, which uses the LM375, an IC designed especially for oscillator applications, comes straight from the National Semiconductor *Linear IC Data Book*. I've found its performance to be very good. The application shown here is for a TTL-compatible output. A sine wave output is obtained by deleting the TTL buffer and feeding the oscillator output into the linear input (pin 3) of the buffer, bypassing the limiting input (pin 2) with a 0.01 μF capacitor. Harmonic output is quite low.



L7	C7	mode	C2
50 kHz 2 mH	0.01 µF	series	1000 pF
100 kHz 1 mH	2200 pF	parallel	2/3 specified load capacitance (i.e., 30, 50, 100 pF)

fig. 8. Low-frequency oscillator with tuned circuit (after Foster and Rankin, reference 7). Recommended for GT-cut (surplus) crystals in the 40 to 100 kHz range or for hard-to-start DT-cut crystals. Series- or parallel-mode crystals may be used. Q1 and Q2 are 2N2979, 2N3565, 2N5770, BC107, BC109, etc.

This is a parallel-mode oscillator. Load isolation with the buffer is extremely good; wide variations in load have negligible effect on frequency. Note the low output impedance of the buffer. The circuit constants given work with crystals between 800 kHz and 3 MHz. Output voltage depends on the power-supply voltage. When using the buffer in the linear mode, a supply of up to 24 volts can be used.

high-frequency oscillators

Modern crystals manufactured for fundamental operation between 3 and 20 MHz are generally AT cut, whereas the older FT243 crystals were generally BT cut. The AT-cut crystal has about an order-of-magnitude better temperature stability performance than the BT cut, but the latter produces thicker crystal blanks, which are far more robust, particularly at the upper end of the range. However, an oven-controlled BT-cut crystal has stability somewhat better than an AT cut crystal, even in an inexpensive commercial oven (see "Oscillator Topics" in reference 11).

The first circuit (fig. 13A) is described by Lane[6] and has the advantage of not requiring a tuned circuit. It is, of course, the same circuit as in fig. 6 but adapted to the high-frequency range. This circuit is recommended for noncritical applications where high harmonic output is desired. It particularly suits FT243 crystals, which may have a high *esr*. Stability is improved and harmonic level reduced when resistor R_f is increased. However, as noted previously, the circuit takes longer to reach full output.

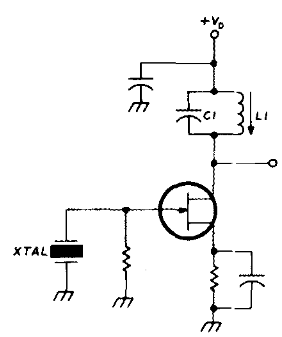


fig. 9. Fet crystal oscillator often seen in the literature. Circuit suffers from frequency pulling with load changes. Performance repeatability is questionable.

The addition of a tuned circuit, as shown in fig. 13B, considerably reduces harmonic output. The tuned circuit should have a high unloaded Q for best results. With an unloaded Q of 50 and R_f of 3.9k, the second harmonic was 35 dB down. When a coil having an unloaded Q of 160 was substituted (R_f still 3.9k), the second harmonic decreased to −50 dB. Output voltage also increases with a high-Q coil. A link winding on the coil, or a capacitor from the transistor collector, can be used for output coupling. Tuning the coil pulls the frequency, but this adjustment is generally made only once and pulling is thus of little consequence. An output buffer is recommended to reduce frequency pulling with load variation. The circuit is for parallel mode operation. A temperature stability of ±10 ppm, largely dependent on the crystal, is readily achieved over the temperature range 32° to 140° F (0° to 60° C).

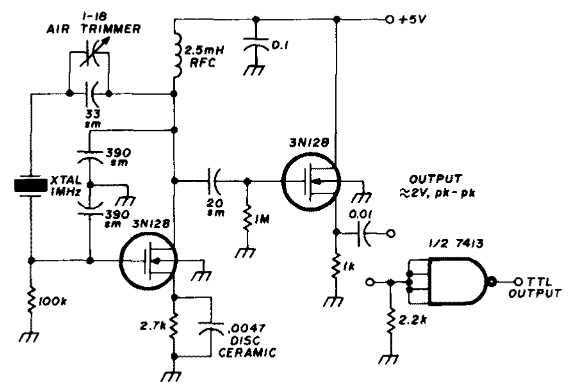


fig. 10. Recommended circuit for a highly stable 1-MHz source. With an inexpensive oven, stability approaching 1 part in 10^{-8} per day is achievable.

Another simple circuit that doesn't require a tuned circuit appears in fig. 14A. This circuit is described by Foster and Rankin,[7] and variations are common in the literature. Some use an rf choke or tuned circuit in place of R1, although this isn't necessary. More output can be obtained by putting a 1.5k load resistor in the collector of Q1 and taking output from the collector, as shown in fig. 14B. About twice the output level may be obtained than in fig. 14A. A TTL output can be obtained using the combination of fig. 14B and C. However, sufficient drive for a counter or other TTL device may be difficult to obtain at the upper end of the frequency range.

In contrast to the untuned version of fig. 13A, the harmonic output of fig. 14A is quite low. It depends on the transistor and the power-supply voltage, but the harmonic level is generally better than −30 dB. Frequency

stability is affected about 0.001% with supply variation between 5 and 10 volts. A temperature stability of +10 ppm can be expected, as for fig. 13A, over the range 32° to 140°F (0° to 60°C).

The circuit in fig. 15 is suitable for crystals specified for operation in the series mode. Coil L1 is adjusted to set the frequency. Performance is similar to the circuit of fig. 14A. This circuit is otherwise known as the "impedance-inverting" circuit. To set it up and to ensure that the coil resonates near the crystal frequency, apply a short circuit across the crystal terminals and tune the coil until the output is close to frequency, then remove the short and tune L1 to set the correct frequency.

Under *no* circumstances should a tuned circuit be placed in the Q1 collector that resonates to the crystal frequency. The circuit may then oscillate as a tuned-base, tuned-collector oscillator not under control of the crystal. A tuned circuit on a *multiple* of the crystal frequency can be used for frequency multiplication, where

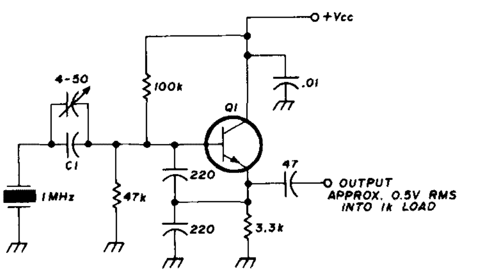


fig. 11. Parallel-mode 1-MHz oscillator usable with crystals between 800 kHz to 3 MHz. Output is a sine wave at about 0.5 volt rms into a 1k load. Note that C1 differs for different crystal-load capacitances. Q1 is a 2N3563, 2N3564, 2N3693, 2N5770, BC107, BC547 or SE1010.

crystal load (pF)	C1 (pF)
50	68
30	remove

a minimum number of components and frequency multiplication are required. Take the output from the collector circuit in such case. (These comments also apply to fig. 14A.)

The oscillator in fig. 16A is discussed in detail by Lane.[6] It's a series-mode oscillator, but parallel-mode crystals can be used, as discussed earlier. Adjust feedback by varying the ratio of C1 and C2. Tune the slug in L1 for frequency trimming.

Resonating L1 is quite simple. Using a grid-dip oscillator resonate L1 with C1 in parallel. In-circuit adjustment can be made with the grid-dip oscillator. Short circuit the crystal and dip the inductor to frequency. Check circuit oscillation by shorting the crystal and applying supply voltage. Tune L1 until output is close to the crystal frequency, then remove the short and tune the slug in L1 to pull the crystal exactly onto frequency. A buffer is recommended as the frequency is sensitive to load variations. Less variation occurs with supply voltage variation.

Crystal power dissipation can be varied by adjusting R1, whose value should be between 100 and 1000 ohms.

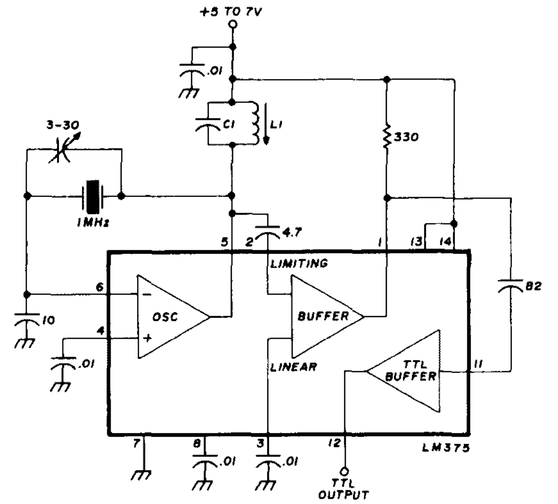


fig. 12. A 1-MHz IC oscillator with TTL-compatible output. Performance is good; harmonic output low. Circuit constants are for crystals between 800 kHz to 3 MHz. C1-L1 resonates at the crystal frequency; if L1 = 0.5 mH rfc, C1 = 47 pF in parallel with 30 pF trimmer.

The lower the value, the lower the crystal power dissipation and the better the stability. Output will drop a little with lower values of R1. A temperature stability of 10 ppm can be obtained with this circuit.

Fet fundamental-frequency crystal oscillators derive directly from vacuum-tube technology. Figs. 17, 18 and 19 illustrate the use of fets. When using mosfets, gate bias is obtained by the diode from gate to ground, as in figs. 18 and 19. Figs. 17 and 18 are the familiar Colpitts oscillator. Harmonic output is low and depends on the fet used. Crystal power dissipation is reduced by reducing C1 capacitance.

Fig. 19 is a Miller oscillator that seems to provide improved performance over most circuits of this type. The circuit also illustrates the single-stage oscillator-

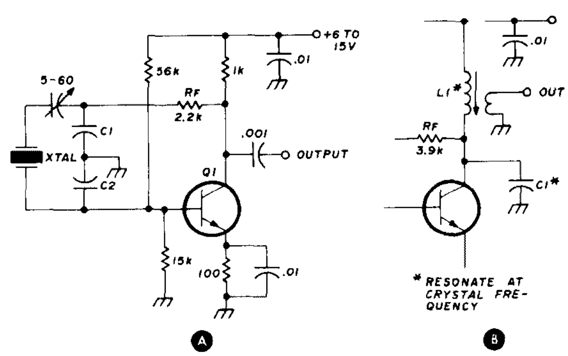


fig. 13. Fundamental-frequency, parallel-mode oscillator for 1 to 18 MHz (A). A buffer is recommended. Adding a tuned circuit (B) reduces harmonic output. Q1 is a 2N916, 2N918, 2N3564, 2N5770, BC107, BC108, BC109, BC162, BF180, BF200, SE1001 or SE1002.

crystal freq (MHz)	C1 (pF)	C2 (pF)
1-3	470	820
3-10	220	470
10-18	120	330

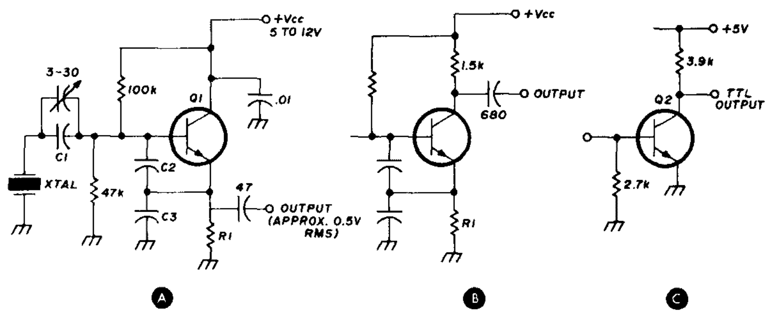
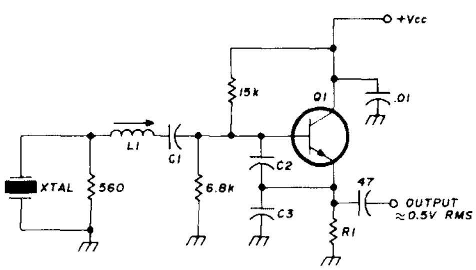


fig. 14. Another version of a fundamental-frequency, parallel-mode high-frequency oscillator requiring no tuned circuit (A). Circuit (B) delivers about twice the output than (A). TTL output can be obtained using the combination of (B) and (C). Q1 is a 2N3563, 2N3564, 2N3693, 2N5770, BC107, BC547, BF180 or BF200; Q2 is a 2N3478, 2N3932, 2N3933, 2N4259 or 2N5179. Other component values are shown at bottom right.

crystal freq (MHz)	C2 (pF)	C3 (pF)	R1 (ohms)	C1
3-10	150	150	2.7k	120 pF for 50 pF specified load capacitance; 39 pf for 30 pF specified load capacitance
10-20	82	82	1k	none for 20 pF specified load capacitance

multiplier application. A tuned circuit resonating at a multiple of the crystal frequency can also be placed in the oscillator drain circuits of figs. **18** and **19**. A temperature stability of 15 to 20 ppm can be expected from these circuits over the operating temperature range of the crystal (assuming use of the AT or BT cut). Parasitics can be troublesome with these circuits, as most fets have good gain well into the vhf range. A **10**- to 47-ohm resistor in series with the gate lead, as shown in fig. **17**, usually overcomes the problem.



crystal frequency	C1	C2, C3	R1	L*
3-10 MHz	1000 pF	270 pF	1500 ohms	2-4 MHz, 60 turns; 4-6 MHz, 40 turns; 6-10 MHz, 25 turns
10-15 MHz	100 pF	220 pF	680 ohms	15 turns
15-20 MHz	100 pF	100 pF	680 ohms	10 turns

*All inductors closewound with no. 33 (0.2mm) enamelled wire on 1/4" (6.5mm) diameter slug-tuned coil form.

fig. 15. High-frequency oscillator circuit suitable for crystals specified for series-mode operation. Known as the "impedance inverting" circuit, its performance is similar to the circuit of fig. 14A. Q1 is a 2N3563, 2N3564, 2N3963, 2N5770, BC107, BC547, BF180, BF200 or SE1010.

Fig. **20** is a high-frequency version of the circuit in fig. **12**, using the LM375 IC. Output voltage depends on the supply voltage. *The circuit will oscillate with a* supply voltage down to 4 volts. Tuning L1 will pull the crystal frequency, but this adjustment should not be used to trim the frequency. Tune L1 only for maximum output. If C3 is a 3 to 30 pF trimmer, it should be used to set the crystal to frequency. Again, harmonic output is low. Using a tuned circuit in the buffer output is likely to cause the circuit to oscillate of its own accord, uncontrolled by the crystal. Use a high-Q tuned circuit for L1/C1 if possible. Stability approaches the crystal specifications; however, the circuit is sensitive to supply-voltage variations.

The buffer provides excellent load isolation and has a low output impedance. **Fig. 21** is the series-mode version of the circuit in **fig. 20.** In this case L1 can be used to trim the crystal frequency. Similar comments apply for the other characteristics.

TTL IC oscillators

I've made brief mention of obtaining TTL-compatible outputs from crystal oscillators to drive digital circuitry (under *1 MHz oscillators*). If the output level of an oscillator is sufficient, around 2 volts peak, a 7413 Schmitt trigger makes a good TTL-output buffer, as in **fig. 10.**

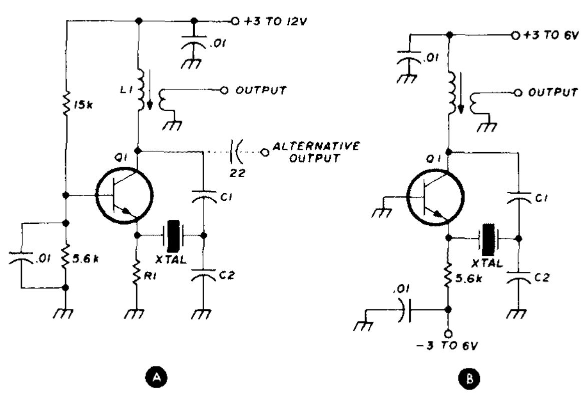


fig. 16. High-frequency crystal oscillator for series- or parallel-mode operation (A). A minimum-component version is shown in (B). The lower the value of R1, the lower will be crystal power dissipation and the better the stability. Q1 is a 2N918, 2N3564, 2N5770, BC107, BC108, BF115, BF180, BF200, SE1001 or equivalent. L1 resonates to crystal frequency with C1.

crystal freq (MHz)	C1 (pF)	C2 (pF)
3-10	47	390
10-20	22	220

Another solution is illustrated in **fig. 14**, where oscillator output may be below **1** volt (but it must be at least **0.6** volt peak).

The combination of a discrete oscillator followed by a TTL buffer may, however, present problems where space is limited, although the circuits in **fig. 14** can be made fairly compact. The **LM375 IC** provides one solution, although a tuned circuit must be used. For low cost and convenience or where space is at a premium, the common TTL gates can be pressed into service and, indeed, are in common use. **Fig. 22** illustrates a common method of using a 7400 NAND gate as a crystal oscillator. Two gates are biased into their linear region and coupled as an oscillator with the crystal in the feedback path. A third gate is used as a buffer.

These three circuits (**figs. 22, 23** and **24**) are aperiodic and would certainly have applications where a high output, aperiodic oscillator is called for. They make excellent frequency markers. Temperature stability is somewhat worse than that of discrete circuits. If you need a very stable oscillator in your counter, you'll just have to make room for more circuitry.

The circuit of **fig. 22** will accept crystals from 1 to 10 MHz. Some trouble may be experienced with crystals above this frequency. Occasionally, trouble may be had with unreliable starting performance, which is largely a function of the two bias resistors. Change their value up or down by using the nearest preferred value to see what effect it has. A wide variation should not be necessary. Frequency and starting performance are critically dependent on supply voltage.

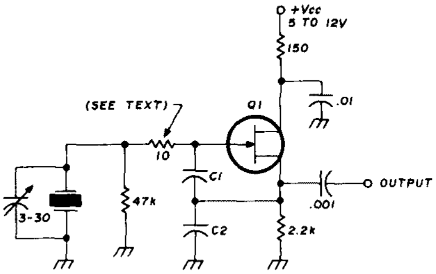
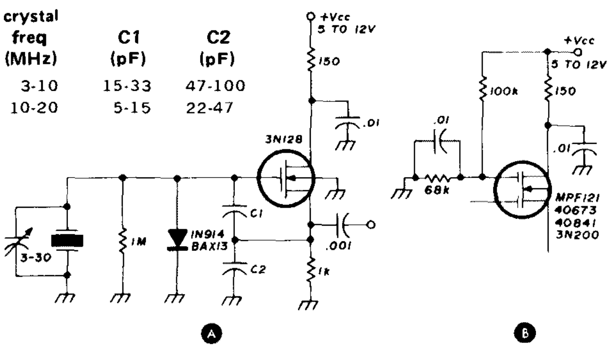


fig. 17. Fundamental-frequency, parallel-mode oscillator using a jfet. Familiar Colpitts circuit is used. Q1 is a 2N3819, 2N5245, 2N5459, 2N5485, MPF102, or MPF104-106.

crystal freq (MHz)	C1 (pF)	C2 (pF)
3-10	27	68
10-20	10	27



crystal freq (MHz)	C1 (pF)	C2 (pF)
3-10	15-33	47-100
10-20	5-15	22-47

fig. 18. Igfet fundamental-frequency, parallel-mode oscillator using single-gate (A) and dual-gate (B) transistors. As in the circuit of fig. 17, harmonic output is low and depends on the device used.

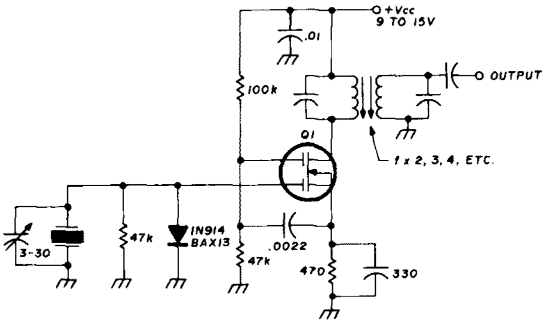


fig. 19. Fundamental-frequency, parallel-mode Miller oscillator illustrating single-stage oscillator-multiplier application. A tuned circuit resonating at a multiple of the crystal frequency can be placed in the drain circuit in this arrangement and that of fig. 18. Q1 is a 3N200, 2N209, 3N210, 40673, 40841, or MPF 121.

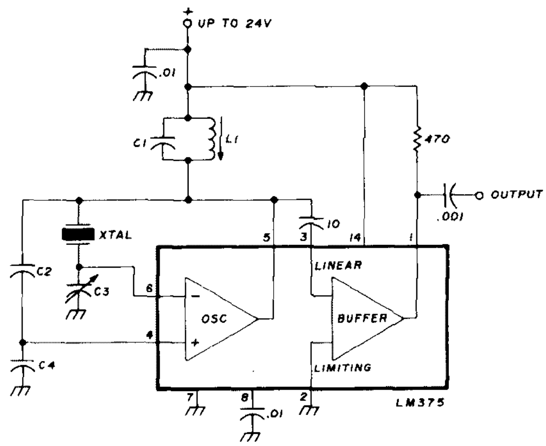


fig. 20. A high-frequency version of the circuit of fig. 12, using the LM375 integrated circuit. L1-C1 is resonant at the crystal frequency. Coil L1 should be tuned for maximum output only, not to trim the crystal frequency. A tuned circuit is not recommended in the buffer output to avoid self-oscillation.

crystal freq (MHz)	C2/C3 (pF)	C4 (pF)
3-10	22	180
10-20	10	82

Where it may be inconvenient to use a NAND gate, a NOR-gate oscillator can be substituted, using a 7402 or other TTL type, as shown in **fig. 23**. This circuit has performance similar to the circuit of **fig. 22**. Here, also, trouble may be experienced with starting: vary the gate bias resistors, as mentioned above.

The circuit in **fig. 24** overcomes problems with poor

starting performance and has a higher upper-frequency limit than those of **figs. 22** and **23**, approaching the limit of the NAND gate used. I've used this circuit successfully with crystals over the range 1 to 20 MHz and have experienced no trouble. It starts reliably every time, even with crystals having a high *esr*. I thoroughly recommend it. The circuit comes from K1PLP[12] and is after Weggeman.[13] For a good discussion on TTL oscillators I recommend both references.

When constructing oscillators using TTL ICs it's wise to bypass the power supply pin as close to the IC pin as possible. Keep the ground lead short, too.

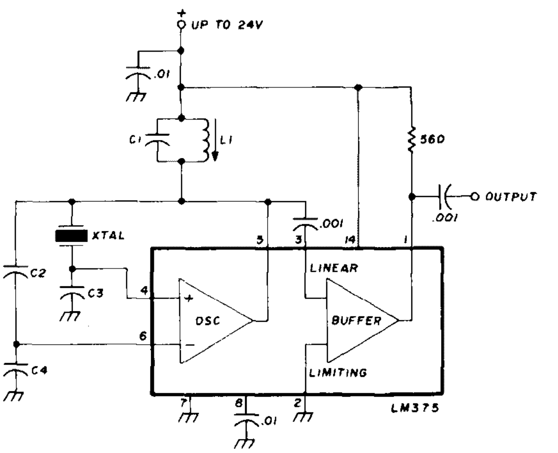


fig. 21. Series-mode version of the circuit in fig. 20. L1-C1 is resonant at the crystal frequency; L1 can be used to trim the crystal frequency in this case.

crystal freq (MHz)	C2 (pF)	C3 (pF)	C4 (pF)
3-10	68	68	330
10-20	27	27	120

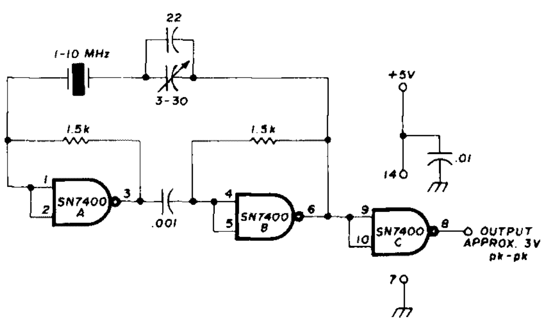


fig. 22. TTL NAND-gate crystal oscillator useful for applications requiring a high-output aperiodic oscillator. Crystals between 1 to 10 MHz can be used.

overtone oscillators

Modern crystals manufactured for overtone operation are usually AT cut, although the BT cut is sometimes found. Take heed of the comments on overtone crystals in the introductory paragraphs of this article. All overtone circuits operate with the crystal in the series mode. Overtone crystals are calibrated at the overtone frequency, not the apparent fundamental.

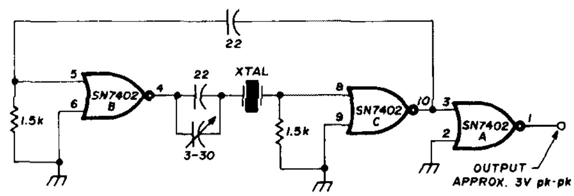


fig. 23. NOR-gate version of the circuit in fig. 22. Difficult crystal starting may occur in these TTL gate circuits, in which case the gate bias should be varied.

An overtone oscillator should not require excessive drive for the crystal to oscillate on the overtone frequency and should have some provision separate from other circuit functions for setting the frequency. In addition, it should not be possible for the crystal to oscillate on subharmonics of the overtone frequency, or on the fundamental. Reliable starting is also desirable.

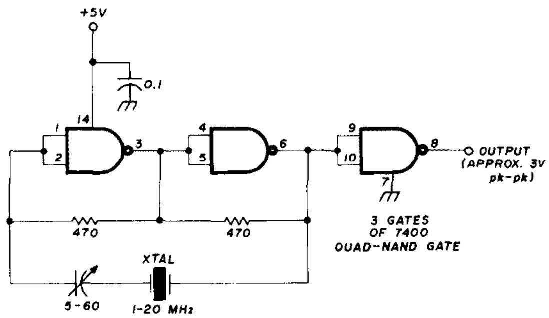


fig. 24. TTL crystal oscillator (after Weggeman, reference 13). Circuit overcomes poor crystal-starting problems and has a higher upper-frequency limit than the circuits of figs. 22 and 23.

The Miller oscillator, **fig. 25**, using an fet, has the advantage of requiring minimum components but doesn't give reliable starting performance. L1 can be

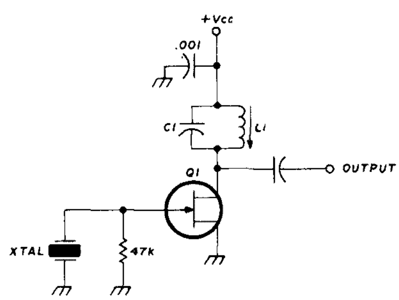


fig. 25. Miller overtone oscillator circuit. Not recommended because of problems with frequency instability due to output load variations and unreliable crystal-starting performance. L1-C1 is tuned to the overtone frequency. Q1 is a 2N3819, 2N5245, 2N5485, MPF102, MPF104-106, or T1S88.

tuned to pull the crystal frequency but may not give enough range; output varies also. Apart from these difficulties, it's prone to oscillating on the fundamental or subharmonics of the overtone. Output load variations adversely affect the frequency. A simple circuit but not really recommended, as performance cannot be guaranteed.

The modified Colpitts overtone oscillator in **fig. 26** is commonly encountered. Unlike the Miller circuit, reliable starting can be expected. It's suitable only for third-overtone crystals, as fifth-overtone crystals exhibit a much higher *esr*. Some third-overtone crystals having a high *esr* sometimes give trouble. Tuning L1 or varying the emitter resistor will pull the frequency, but output will vary. Feedback can be varied by changing C2. Harmonic output depends on the transistor used and the supply voltage. This circuit has the advantage of simplicity but is not recommended.

Fig. 27 illustrates a Pierce-type overtone oscillator. Note that a single-gate fet may be used. Crystal starting is sometimes unreliable and excess feedback may be

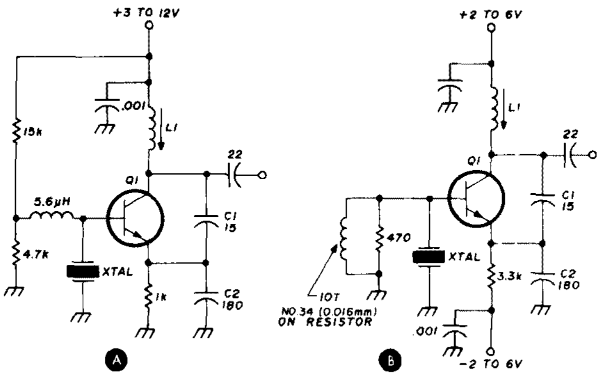


fig. 26. Third-overtone crystal oscillator for 20 to 60 MHz (A) and a minimum-component version (B). Although commonly seen in the literature, it's not recommended for reasons discussed in the text. L1-C1 is resonant at the overtone frequency. Q1 is a 2N3563, 2N3564, 2N5770, BF180, BF200 or SE1010.

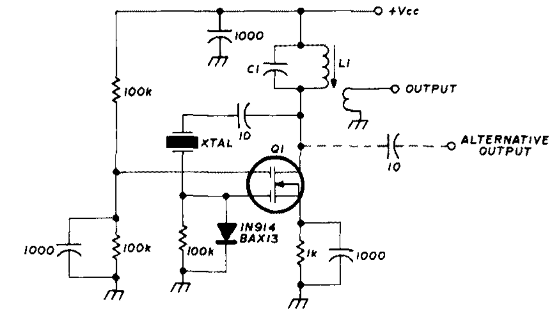


fig. 27. Pierce-type overtone oscillator for 15 to 60 MHz. Not recommended for the same reasons discussed in connection with the circuits of figs. 25 and 26 (see text). L1-C1 resonates to the crystal frequency. Q1 is a 3N200, 3N209, 40673, 40841, BFS28, MPF121, MPF131, or equivalent.

required. No provision is made for setting frequency, but as in the previous two circuits, tuning L1 pulls the crystal. The comments on the circuits of **figs. 25** and **26** also apply here.

When overtone crystal-oscillator circuits first appeared in the literature, they had a tendency to be somewhat complicated to avoid the problems discussed above. The following circuits are uncomplicated and reliable.

The circuit of **fig. 28** provides all the desired characteristics with a minimum of components. Starting is reliable and crystal power dissipation is well below the allowable maximum. Increasing R1 increases crystal power dissipation. Stability is as good as that of a fundamental-frequency oscillator. This circuit is discussed in detail by Lane.[6]

Coil L1 can be roughly set to frequency with no supply voltage by shorting the crystal and dipping L1 with a grid-dip oscillator. Apply power, and while monitoring the output frequency, tune L1 close to the marked crystal frequency. Remove the short and trim the frequency with the 3 to 30 pF trimmer. Tuning L1 will pull the crystal, but L1 does not require tuning following the initial setup. Harmonic output is low, usually

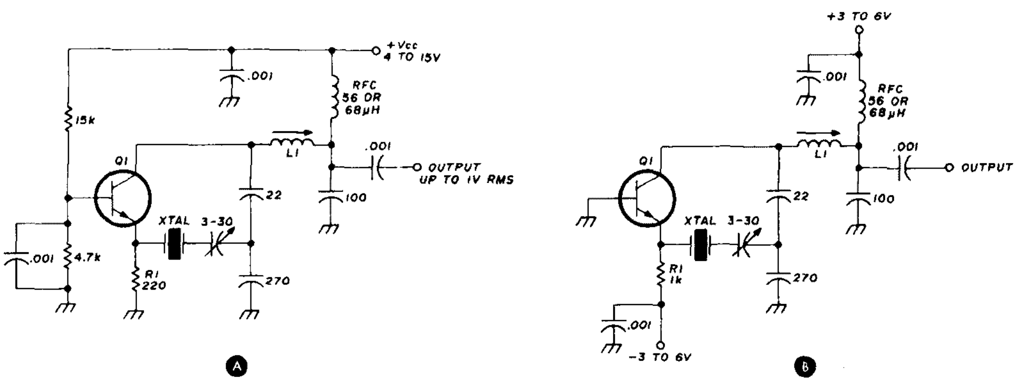


fig. 28. Third-overtone oscillator for 15 to 65 MHz (A) and a minimum-component version (B). All the desirable characteristics for this application are provided in this circuit. Crystal starting is reliable and crystal power dissipation is well below the allowable maximum. Q1 is a 2N818, 2N3563, 2N3564, 2N5770, BF180 or BF200. L1 resonates at the crystal frequency with 22 pF (1 μH for 15 to 30 MHz; 0.5 μH for 30 to 65 MHz).

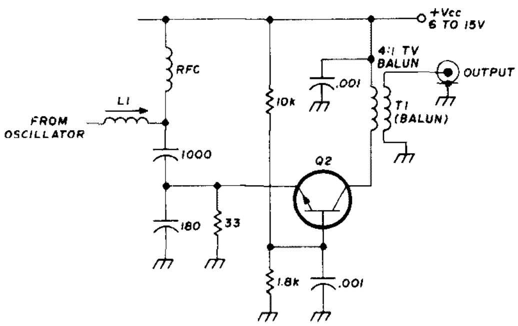


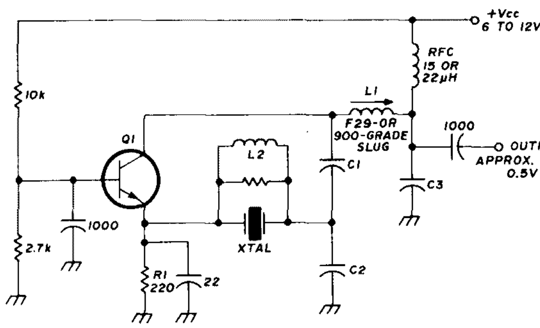
fig. 29. Buffer amplifier for the circuits of figs. 28 and 31. Circuit provides excellent performance and supplies several volts of output. Input and output tuned circuits must be well shielded. T1 is a 4:1 TV-type balun. Q2 is a 2N3563, 2N3564, 2N3646, 2N5770, BF180, or BF200.

around −40 dB, but depends on the transistor and the supply voltage.

A buffer is recommended. The circuit of **fig. 29** is excellent, providing several volts of output. Output transformer T1 can be a standard **300**- to 70-ohm vhf television balun or a homebrew balun. Alternatively, a single or double-tuned circuit can be used. L1 and the output tuned circuits must be well shielded and separated, otherwise instability may result. Q2 bias may need to be varied, depending on the transistor. Tune L1 for maximum output before trimming the crystal frequency. For multichannel operation, switching the crystal and trimmer is permissible, but leads must be kept very short.

The "impedance-inverting" circuit discussed by Foster and Rankin[7] also has much to recommend it. This circuit is illustrated, for third-overtone crystals, in **fig. 30**. Performance is similar to the circuit of **fig. 28**. Coil L1 is used to trim the crystal frequency. The 560-ohm resistor across the crystal prevents oscillation on modes other than the overtone. This circuit readily adapts to a single-stage oscillator multiplier as we shall see later. It is more amenable to multichannel operation, as one side of the crystal is grounded. Note that if L1 is made too large oscillation on frequencies other than the overtone may occur.

Overtone crystals above **65** MHz are usually fifth- or seventh-overtone types. The circuit in **fig. 31** is from Lane[6] and is a variation of the circuit in **fig. 28**. The rf



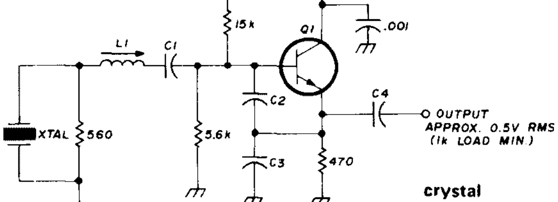
L1	65-85 MHz:	7 turns no. 22 (0.6mm) or no. 24 (0.5mm) enamelled, closewound on 3/16" (5mm) diameter form
	85-110 MHz:	4 turns no. 22 (0.6mm) or no. 24 (0.5mm) enamelled, on 3/16" (5mm) diameter form, turns spaced one wire diameter
L2		10 turns no. 34 (0.2mm) closewound on low-value ¼-watt resistor

C1	65-85 MHz:	15 pF	85-110 MHz:	10 pF
C2	65-85 MHz:	150 pF	85-110 MHz:	100 pF
C3	65-85 MHz:	100 pF	85-110 MHz:	68 pF

fig. 31. Oscillator for fifth- or seventh-overtone crystals in the 65 to 110 MHz range. The buffer circuit of fig. 29 is recommended. Q1 is a 2N3563, 2N3564, 2N5770, BF180, BF200 or SE1010.

choke formed by L2, wound on a low value resistor, suppresses the lower resonances of the crystal, ensuring operation on the correct overtone. The comments on the circuit of **fig. 28** also apply here. The buffer of **fig. 29** is also recommended. The circuit in **fig. 32** is a variation on the "impedance-inverting" oscillator from Foster and Rankin.[7] Performance is similar to the circuit of **fig. 31**. Again, a buffer is recommended. The four circuits just described, **figs. 28, 30, 31** and **32**, are all slightly frequency sensitive to supply voltage variations, so a well-regulated supply is recommended.

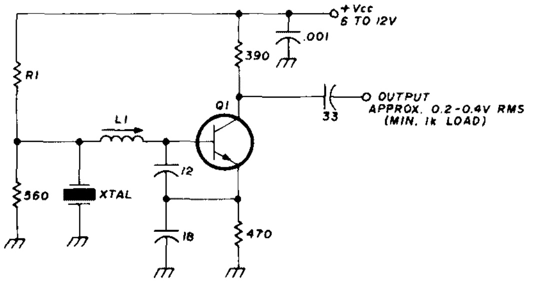
The oscillator of **fig. 30** can be readily adapted to a single-stage oscillator multiplier and has ready application in vhf converters. Good load isolation and low unwanted output levels are afforded by the circuit of **fig. 33**. The double-tuned circuit is the key to success here. Coils L2, L3 should be no more than critically coupled,



crystal freq (MHz)	C1 (pF)	C2 (pF)	C3 (pF)	C4 (pF)	L1 ¼" (6.5mm) diam form	L1 3/16" (5mm) diam form
15-25	100	100	68	33	12 turns no. 30 (0.25mm) closewound	15 turns no. 28 (0.3mm) closewound
25-55	100	68	47	33	8 turns no. 30 (0.25mm) closewound	10 turns no. 28 (0.3mm) closewound
50-65	68	33	15	22	6 turns no. 22 (0.6mm) space to ¼" (6.5mm) long	7 turns no. 28 (0.3mm) closewound

fig. 30. The "impedance inverting" circuit for third-overtone crystals in the 15 to 65 MHz range. Performance is similar to the circuit of fig. 28. Q1 is a 2N3563, 2N3564, 2N5770, BF180, BF200 or SE1010.

which results in the best output for good harmonic rejection. All outputs, other than the multiple of the crystal frequency, are about −60 dB or greater. Tuning L2 and L3 pulls the frequency slightly, but as this is usually a once-only adjustment, it is of little consequence. L2 has a broad tuning characteristic because of Q1 loading,

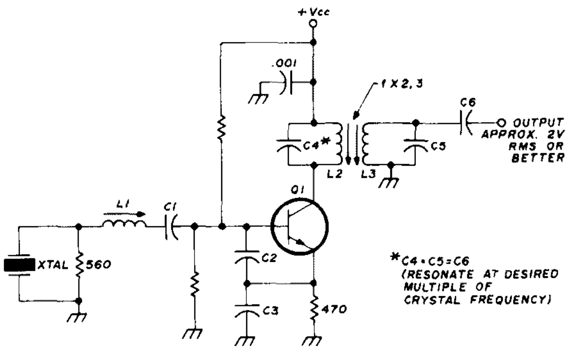


crystal freq (MHz)	L1	R1 (ohms)
60-85	7 turns spaced 1/4" long*	3300
80-110	4 turns spaced 1/4" long*	1800

*Wound on 3/16" (5mm) diameter form with vhf slug, no. 22 (0.6mm) enamelled wire

fig. 32. Another variation of the "impedance inverting" oscillator for 60 to 110 MHz fifth- or seventh-overtone crystals. Performance is similar to the circuit of fig. 30. Q1 is a 2N3563, 2N3564, 2N5770, BF180, BF200, or SE1010.

whereas L3 tunes quite sharply. The circuit is ideally suited to fet mixers having local-oscillator injection into the gate. A word of caution: the connection between Q1 collector and tuned circuit L2/C4 must be kept extremely short, otherwise parasitic oscillations will occur.



L2,L3 60-90 MHz: 9 turns no. 22 (0.6mm) enamelled wire, closewound on 3/16" (5mm) diameter form with vhf slug

90-130 MHz: 6 turns no. 22 (0.6mm) enamelled wire, closewound on 3/16" (5mm) diameter form with vhf slug

Coil forms for L1, L2 are spaced 0.6" (15mm) apart and assembly shielded.

fig. 33. Overtone oscillator-multiplier adapted from the circuit of fig. 30. This version is applicable to vhf converters. It provides good load isolation and low unwanted output levels (refer to fig. 30 for component values).

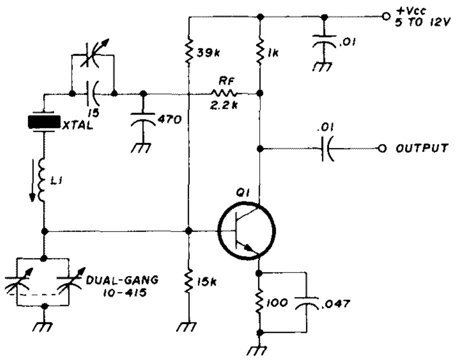


fig. 34. An extended-range vxo circuit (after Lane, reference 6). Maximum frequency shift is nearly 10 kHz at 5 MHz. Circuit is a variation of that in fig. 13, and the same comments apply (see text). Q1 is a 2N3563, 2N3564, 2N5770, BC107, BC547, BF115, BF180, SE1010 or equivalent.

crystal freq (MHz)	L1 (μH)
6-15	5-20
3-6	20-50

variable crystal oscillators

This topic is well covered by Pat Hawker, G3VA,[11] and Lane.[6] A recent article with useful discussions and practical details on this subject is by Doug DeMaw, W1CER;[14] see also reference 15.

The greatest shift you can obtain from AT-cut fundamental-frequency crystals is about 1/500th of the crystal frequency. By far, one of the best circuits I've tried is

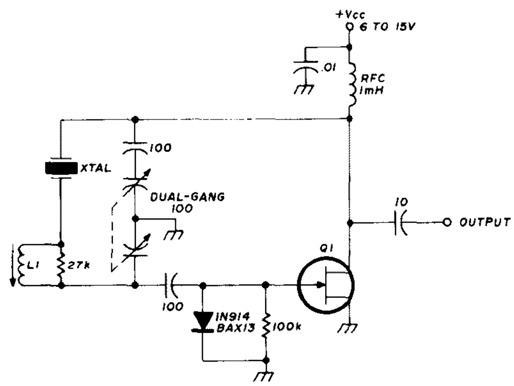
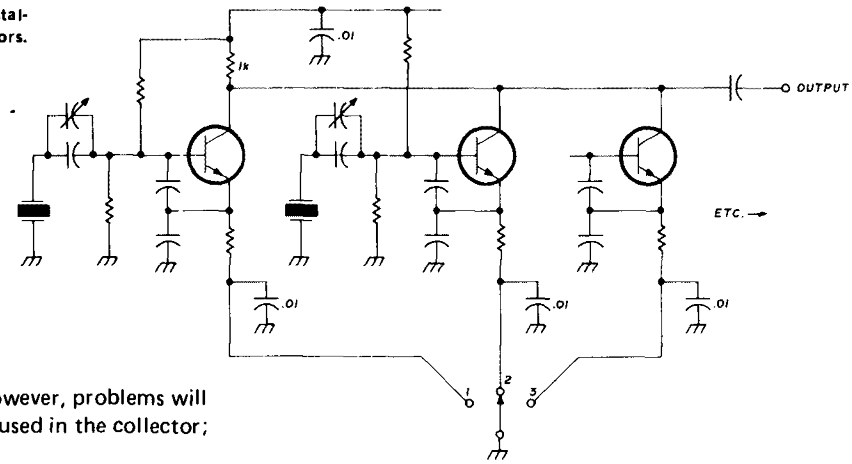


fig. 35. Vxo using a fet. Performance is similar to the circuit of fig. 34. A buffer is recommended for this circuit and that of fig. 34. L1 is a 30 to 80 μH slug-tuned coil. Q1 is a 2N3819, 2N5245, 2N5485, MPF102, MPF104-106, or TIS88.

that of fig. 34. It's described by Lane[6] and is recommended. Frequency stability is not compromised greatly, even at the extremities of the shift. Using a crystal manufactured for vxo operation, the maximum shift obtained using this circuit approaches the maximum obtainable; i.e., nearly 10 kHz at 5 MHz. Harmonic output is high, as this circuit is a variation of that in **fig. 13**. Resistor R_f determines the feedback level and may have to be varied for best starting performance. Other

fig. 37. Another multichannel crystal-switching method using separate oscillators. Outputs are paralleled for dc and ac.



comments on **fig. 13** apply here. However, problems will be experienced if a tuned circuit is used in the collector; the frequency shift is restricted.

The fet vxo in **fig. 35** is commonly seen in the literature; this circuit is after DeMaw.[14] He claims a similar shift to the circuit just discussed. In each case a buffer is recommended. (DeMaw also describes how to drive a tube grid from the buffer.)

No two crystals will be shifted by the same amount; some provide more swing than others. Crystals especially manufactured for vxo operation are obtainable from some crystal manufacturers. Modern crystals perform well; FT243 crystals are generally poor. Those contemplating construction should read references 14 and 15.

Vxos should be built much the same as vfos to achieve best results. Don't use chassis-mounted crystal sockets. Mount the socket on insulating material to avoid stray capacitance. Better still, order crystals with flying leads rather than pins, and solder them directly into the circuit.

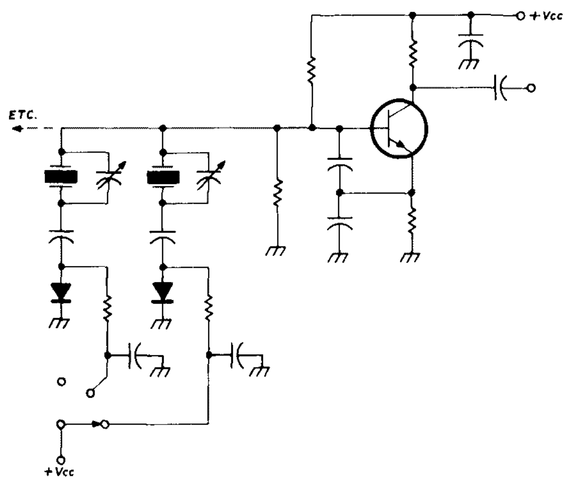


fig. 36. Diode-switching of a multichannel oscillator. Each crystal should be trimmed to frequency individually.

multichannel oscillators

In some applications, switching crystals to change frequency presents practical difficulties. Switching a dc circuit to make a frequency change is better and more convenient practice. Diode switching is one method (**fig. 36**). Each crystal should be individually trimmed. Difficulties do arise, as some crystals and oscillator circuits object to this method.

Another method is illustrated in **fig. 37**. Essentially it's a number of separate oscillators, each switched on in turn. Outputs are paralleled, both for dc and ac. If you want to switch crystals, circuits with one side of the crystal grounded are preferred.

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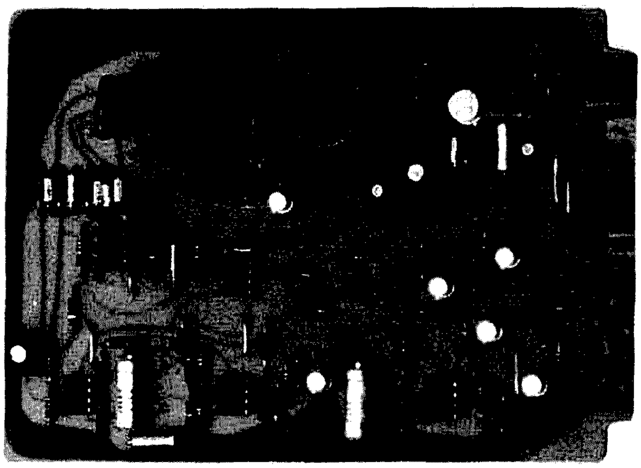

DT-500 RTTY demodulator

Designed with the vhf operator in mind, the DT-500 provides all essential demodulator features at low cost

The DT-600 RTTY demodulator described in an earlier issue of *ham radio*[1] was designed to provide the best possible performance, on the high-frequency bands, that could be obtained at reasonable cost. However, for less critical applications, some of the DT-600 features may be eliminated, with a substantial cost reduction. DT-600 performance isn't required on the vhf bands, for example, nor is its high performance required by the occasional RTTY operator who is on a tight budget.

The DT-500 will provide all the features normally required by the vhf operator and, with the exception of the DT-600, ST-6,[2] and TT/L,[3] will outperform almost all RTTY demodulators now in use on the high-frequency bands.

The CATC systems approach to demodulator design, which was described in reference 1, was applied to the DT-500. Thus the unit is easy to build, simple to interface with other equipment, and is relatively inexpensive.

the DT-500

The circuits in the DT-500 are identical to those in the DT-600. The only differences between the two demodulators are those circuits in the DT-600 that were not included in the DT-500. (The reader is referred to reference 1 for circuit descriptions and adjustment procedures). The input bandpass filter, keying lowpass filter, ATC, and antispace circuits were included in the DT-600 to contend with peculiarities encountered on the high-frequency bands. Adjacent channel interference, noise, selective fading, and CW interference are not usual problems on the vhf bands; hence appropriate circuits for these problems were not included in the DT-500.

Audio from the station receiver is introduced into the DT-500 at a 500-ohm input (fig. 1). The 1000 ohm resistor, R2, is a current limiter, and the two silicon diodes protect the limiter stage from excessive audio voltage. All other parts of the circuit are identical to the corresponding circuits in the DT-600.

By Robert C. Clark, K9HVW, 930 Chestwood Avenue, Tallahassee, Florida 32303

The power supplies (±12 volts and +170 volts) shown in reference 1 are also recommended for the DT-500. Several DT-500s and DT-600s may be operated from the same ±12-volt supply for increased cost effectiveness, but the regulator transistors should be replaced with IC regulators such as the LM 320-12 and the LM 340-12 to handle the required current.

The DT-500 can also function as an extremely low-cost beginner's model. Motor control and mark-hold features may be eliminated by omitting op amps U3, U4, U5, and Q5, Q6. These features account for more than 50% of the components in the unit. If you wish to eliminate these functions for now, they may be added later as operating convenience and budget dictate. If the

interfaces

As mentioned in reference 1, one of the prime concerns of the CATC-designed units is ease of interface with external equipment. Both the DT-500 and DT-600 provide open-collector outputs for data and control lines to interface with low voltage (particularly TTL) peripherals. It's possible to use the open-collector outputs to drive a machine interface and hence keep the high-voltage loop and keyer within the printer. This approach can increase the ease of interconnecting equipment (all interconnecting cables are at low voltage), and at the same time greatly reduce the noise generated by the high-voltage keyer and radiated by the loop. Two machine interfaces, the DI-50 and the DD-350, are

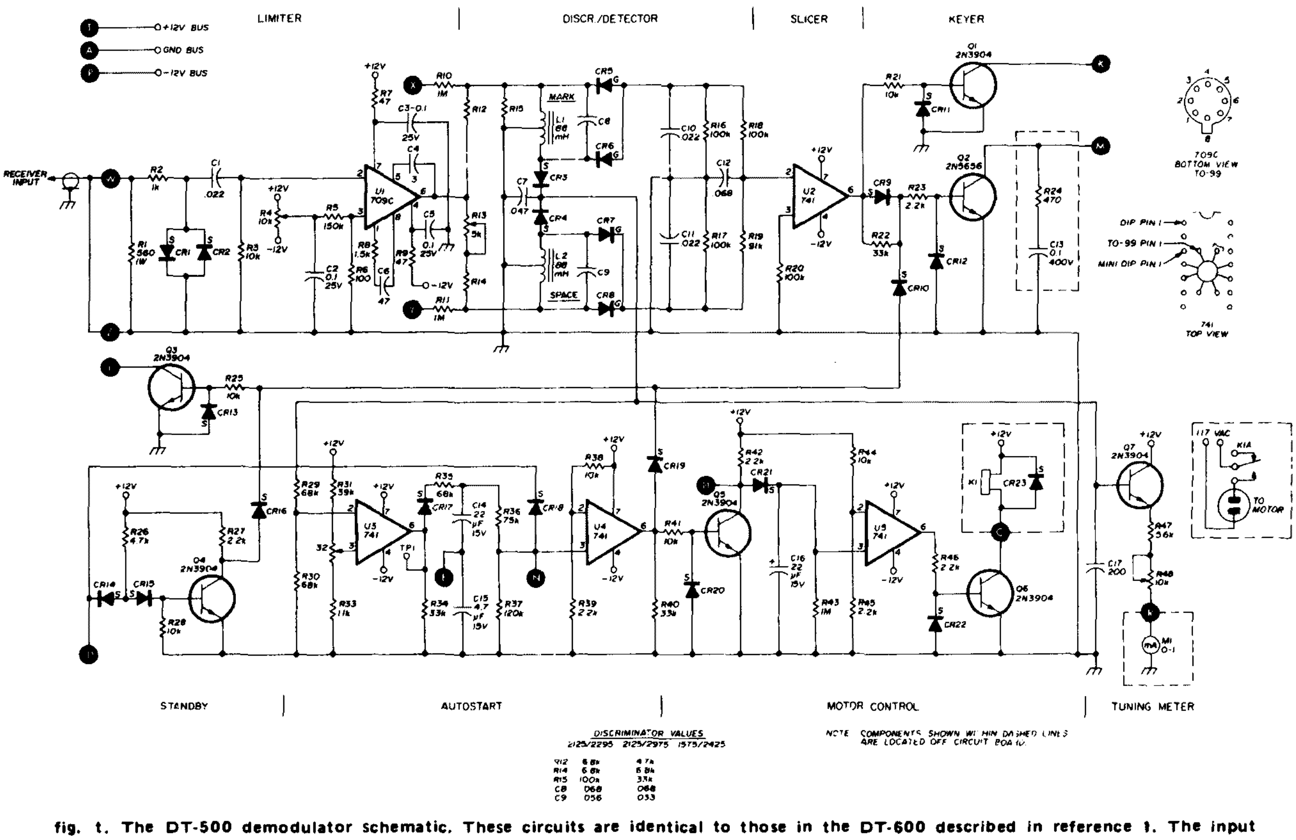

fig. 1. The DT-500 demodulator schematic. These circuits are identical to those in the DT-600 described in reference 1. The input bandpass filter, keying lowpass filter, ATC, and antispace circuits of the DT-600 have been omitted from the DT-500. Diodes marked with a G are germanium 1N270; diodes marked with an S are silicon 50 PIV unless otherwise noted.

bandpass input filter from the DT-600 were added (externally), the DT-500 would outperform the ST-3,[4] ST-4, ST-5[5] and ST-5A and would approach the performance of the ST-6 and DT-600.

As with the DT-600, the DT-500 is built on a single 6x4½-inch (152x114mm) board with a 22-pin edge connector. Fig. 2 shows board interconnections.*

*Double-sided circuit boards with plated-through holes are available from Data Technology Associates, Inc., Post Office Box 1912, Miami, Florida 33143. Send self-addressed, stamped envelope for information.

described here (a number of other interfaces, DI-30 logic-to-polar, DI-40 polar-to-logic, and a loop-to-logic interface have also been prepared).

DI-50 logic-to-loop interface. The DI-50 has its loop keyer floating (no ground in common with the logic portion of the circuitry) and hence may be inserted at any point in the selector-magnet loop. **Fig. 3** is the schematic of the DI-50.

A logic zero at the base of Q4 causes Q4 to conduct

and supply current to the oscillator formed by Q1, Q2. The oscillator operates in the 800-kHz range. The 800-kHz energy is capacitively coupled (and hence decoupled for dc) through C4 and C5 to a bridge rectifier consisting of germanium diodes CR1-CR4. (As shown in fig. 4, the printer loop is in mark when a logic zero exists at the base of Q4 and in space with a logic

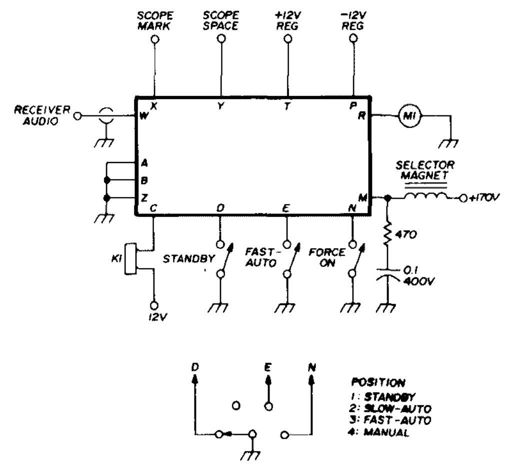

fig. 2. DT-500 demodulator board inter-connections (A) and an alternative circuit for a single switch (B). The unit is constructed on a single 6x4½-inch (152x114mm) circuit board with a 22-pin edge connector.

one.) The rectified output is filtered by C6 and applied to the base-emitter junction of Q3, biasing the transistor into the conducting state. The bridge consisting of CR5-CR8 ensures that the high side of the loop is returned to the collector of Q6 for either the DT-500 or DT-600 through a two-conductor shielded cable, the other conductor connecting the open collector of Q1 (either DT-500 or DT-600) to the DI-50 logic input.

Q1 and Q2 should be a complementary pair such as the 2N3904 and 2N3906. Q4 could also be a 2N3906. Other components are not critical, but the low forward-voltage drop of the 1N270 is essential. Q3 should be a high-voltage transistor (300 volts) capable of 100 mA and 5 to 10 watts dissipation, such as the MPS-A42, 2N5656, or the DTS-702.

DD-350 selector magnet driver with motor control. The DD-350 (fig. 4) requires only a single-conductor shielded cable between the demodulator and the printer to control both the high-voltage keyer and the motor relay. Other features include a choice of input logic (mark = 0 or mark = 1) and open-loop operation during standby to reduce power consumption and heat.

Several optional input configurations may be implemented with U4C and U4D. If the mark output from the demodulator keyer (Q1 of DT-500 and DT-600) is a logic zero, as it is for the DT-500 and DT-600, then it is connected to H (fig. 4). However, data will not pass through U4D unless pin 13 is high. This condition may be accomplished by either connecting pin 13 to the positive supply through a 3.9k resistor, or by jumpering J1 and grounding either E or F. Then, either E or F may function as a disable terminal.

Suppose that the data (mark = 0) is connected to H, and F is grounded, enabling U4D. The data (mark = 0, space = 1) is inverted by U4D. A low (space) at pin 5 of the latch formed by U3A and U3B sets the latch (pin 6 high), which in turn enables U4A. The data passes through U4A and U4B. Pin 6 of U4B is high during mark, reverse-biasing CR1, allowing CR2 to conduct, and forward biasing Q1 (the high-voltage keyer). For space, pin 6 will be low, allowing current to flow through CR1 and removing the forward bias on Q1. At

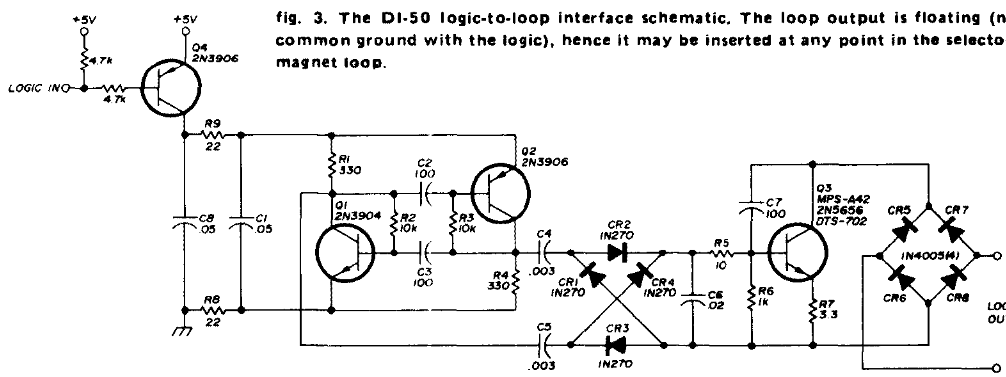

fig. 3. The DI-50 logic-to-loop interface schematic. The loop output is floating (no common ground with the logic), hence it may be inserted at any point in the selector-magnet loop.

connected to the collector of Q3 and that the low side is connected to the emitter, through R7.

If motor control is desired with the DI-50, one side of the coil of a 12-volt relay (K1 in fig. 1) is connected to the 12-volt supply, which in turn powers the 5-volt regulator for the DI-50. The other side of the coil is the time the initial space set the U3A, U3B latch, it also set the latch made up of U3C and U3D and forced Q3 into conduction (discharging C2 and starting the timing cycle of U1). With U3C, U3D set, the high on pin 8 of U3C forward biases Q2, which pulls in the motor relay, starts the timing cycle of U2C, and discharges C5. As

long as the level on the input line continues to alternate, C2 is discharged with every space. If, however, the level remains in the mark condition for 30 seconds, without a transition to space, then C2 charges to the point where U1 triggers. The output pulse (pin 3) goes from high to low, resetting U3C, U3D. As the latch is reset, the motor relay is released, turning off the motor, and the forward DD-350 with only a single-conductor shielded cable connecting the demodulator and the DD-350.

The high-voltage keyer transistor has requirements similar to the transistor described with the DI-50. Note that this keyer stage must be connected in the ground return of the loop supply with the positive side of the loop connected to D.

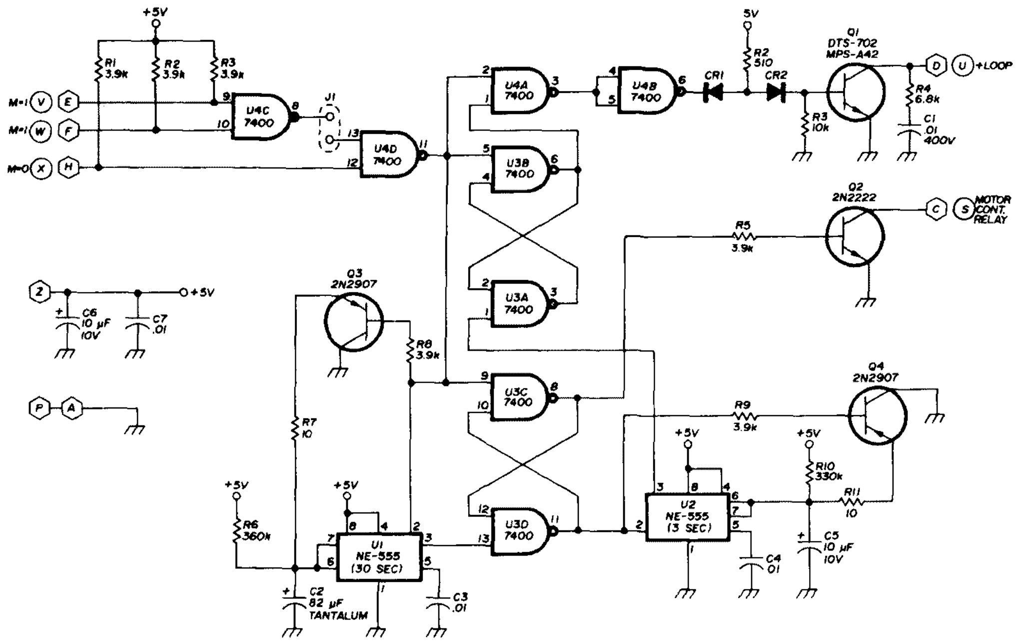

fig. 4. Schematic of the DD-350 selector-magnet driver with motor control. Each PC board contains two circuits as shown. Features include a choice of input logic and open-loop operation during standby.

bias on Q4 is removed, allowing C5 to charge. In three seconds U2 triggers, with the output pulse (pin 3) resetting the U3A, U3B latch and disabling U4A (forcing the loop to space).

Assume the DD-350 has been in standby. When the first character is transferred from the demodulator to the DD-350, the latches are set, the motor starts, and the printer receives the data. As long as the data continues, with less than 30 seconds between characters, the motor continues to run and the characters are printed. When a character has not been received in 30 seconds, the motor shuts down, and three seconds later (to give the machine time to stop), the loop opens to reduce current consumption and heat. The DD-350 will return to the active state immediately with the next character received.

Of course, if Q1 of the DT-500 or DT-600 is connected so that all mark-hold features inhibit the data (jumper C to 2), no data will reach the DD-350 while mark-hold is maintained. The time delay and standby functions of both demodulators prevent data from being transferred (for instance on noise or CW), as does the antispace function of the DT-600. Hence, all features of either demodulator, with the exception of the particular choice of the motor delay, will be preserved by the

the SELCOM

A future article will describe the CATC SELCOM. The SELCOM is a device, based on the universal asynchronous receiver/transmitter (UAR/T), which regenerates distorted RTTY signals, provides for speed changes between the data rate of the RTTY signal and the speed of the printer, and provides a parallel data output for decoding RTTY characters. The decoding and control portions of the SELCOM are also suitable for tone decoders as used in vhf autopatch units. A mini-SELCOM featuring call-letter enable, four-N disable, and two functions for the control of external equipment will also be featured.

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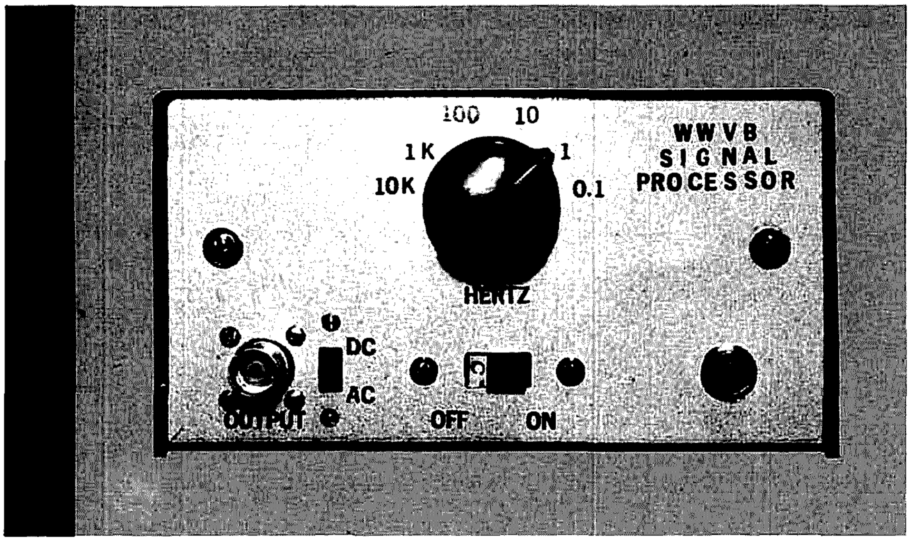


WWVB signal processor

Link your counter's time base to the super accurate signal from the NBS station at Boulder, Colorado

An increasing number of amateurs are using digital counters to determine transmitter frequency in the vhf and uhf region. This technique requires extremely accurate time-base frequencies if the readouts are to be meaningful. The low-cost frequency counters amateurs often use usually don't have accurate time-base crystal circuits; thus obtaining accurate readings with economy counters can be tedious. The time base must be monitored constantly against a primary frequency standard and warm-up time for ovens can be annoying. The answer is not to compare your counter time base with a primary standard but to use the primary standard itself as a time base by linking up with a Bureau of Standards station at Boulder, Colorado.[1]

If WWV is selected, propagation delays will cause timing problems that affect accuracy. Station WWVB is a better choice. Its 60-kHz carrier ensures good accuracy and freedom from shortwave propagation delays. WWVB is the most powerful of the NBS stations (16 kW effective radiated power) and its timing signals are accurate to 2 parts in 10^{11} or better. WWVB's signal received in rural Wisconsin is solid night and day, season to season, except for a slight diurnal effect at local sunrise and sunset. WWVB's signal strength is 500 μV or more throughout the U.S. except for the upper East Coast, lower Florida, and northwestern Washington, where signal strength is 100 μV.

By Harold Isenring, W9BTI, 10850 Amy Belle Road, Colgate, Wisconsin 53017

The following paragraphs explain how to amplify WWVB's 60-kHz carrier, eliminate the nonessential information on that carrier, and use the resultant output in conjunction with your frequency counter to obtain a precision frequency standard.

each second. Further clock information is provided by varying the length of time the carrier is reduced in amplitude each second. The carrier modulation is useless for the project described here and must be removed as it causes an accuracy problem with the final signal.

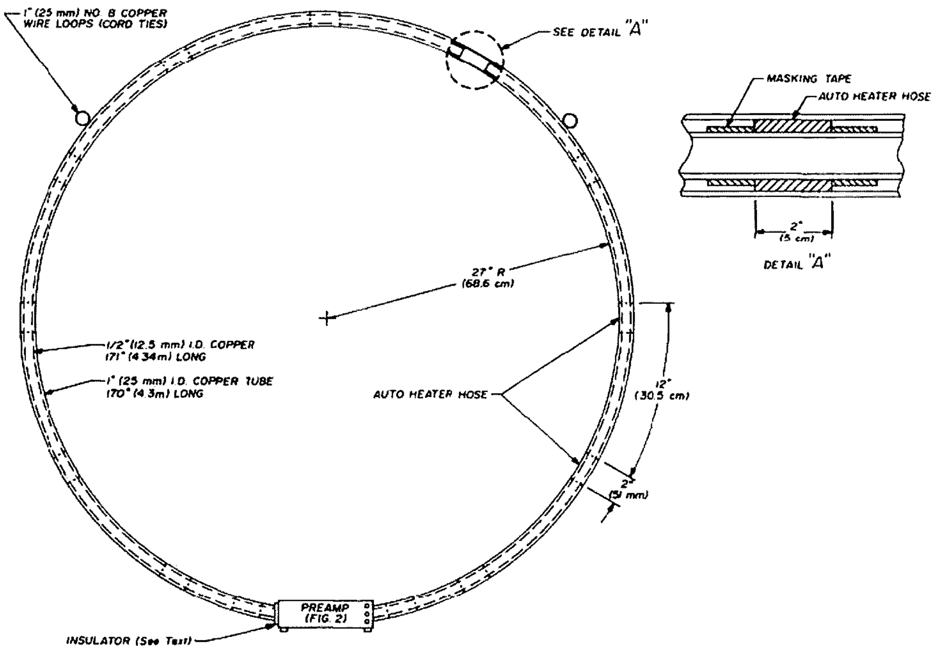


fig. 1. Loop antenna construction details.

preliminary checks

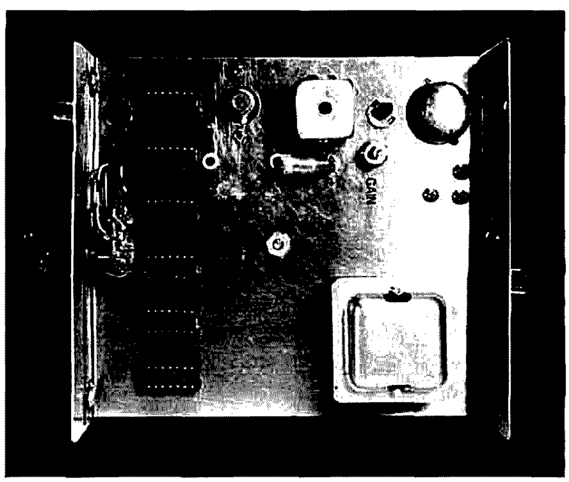
A test setup with receiver, loop antenna and 60-kHz preamp was built to determine the feasibility of this idea. I recommend building the loop and preamp first to determine if WWVB's signal strength is adequate at your location. The dimensions of the loop I built (**fig. 1**) were determined by the size of my attic opening. The large loop size accounted for the great success with WWVB reception. However, my attic location introduced some problems. The attic temperature range was 132°F (70°C), which severely detuned the uncompensated winding within the loop. Severe sensitivity and selectivity losses resulted. The loop now is in my basement, where temperature excursions are only 10°F (6°C) and reception is more than adequate.

The preamplifier (**fig. 2**) can be operated temporarily by connecting it to a 20-volt power supply capable of supplying at least 4 mA. Apply the 20 volts to J1 in series with a 1000-ohm, ½-watt resistor. After amplification by the loop preamp the signal is a pure 60-kHz sine wave and can be observed at test point 1 with an economy oscilloscope.

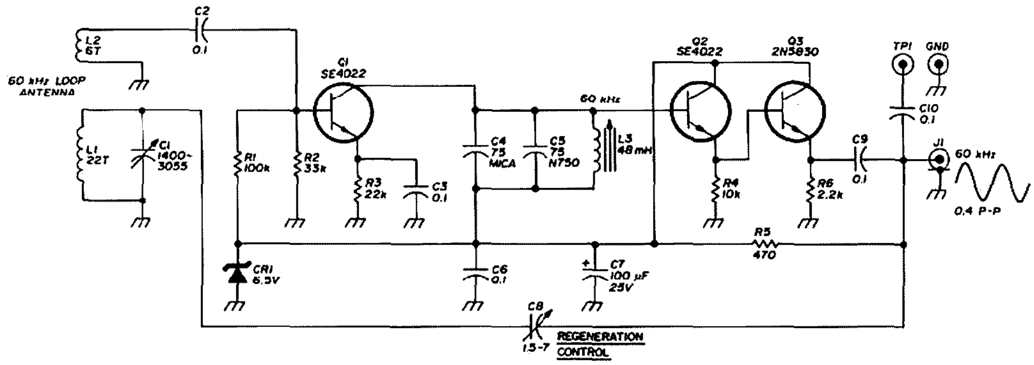
The modulation on WWBV's carrier is capable of controlling a sophisticated digital clock. This modulation is in the form of a 10-dB decrease in carrier strength once

Another problem is caused by the fourth harmonic of the horizontal scanning amplifiers in local television sets. When operating on monochrome reception this signal is 15,750 Hz times four, or 63 kHz. Black and white is rare

Processor top view. Divider chain is immediately behind front panel switch. Rf amplifier, limiter, and zero-crossing detector are at upper left; power transformer is at right.



nowadays, but color broadcasts produce an interference frequency of 15,734.264 Hz times four, or 62.937 kHz. This signal beats with WWVB's carrier resulting in a 2.937-kHz beat note. The beat note is not pure because the interfering signals from television sets are rich in information, a counter will count the frequency as more or less than 60,000. This phenomenon occurs because the squaring circuit turns on and off at a certain point on a sine wave. The squaring circuit turned on at a rising voltage of 0.6 and off at a falling voltage of 0.6. As



L1 22 turns, part of loop antenna (see text)

L2 6 turns, part of loop antenna (see text)

L3 40 kHz remote control cup-core rf coil, 48 mH (found in many TV remote controls)

fig. 2. 60-kHz preamp that is installed in the loop antenna (fig. 1).

harmonics. There are two ways of eliminating this problem: providing 60-kHz receiver selectivity and careful placement of the loop antenna.

WWVB carrier-level shift

The WWVB 60-kHz signal was amplified to a level sufficient to operate a squaring circuit and TTL logic. The WWVB 10-dB carrier reduction, even though still of sufficient amplitude to operate TTL logic, created a new problem. At nearly every 10-dB carrier reduction a frequency shift occurred. If you count the number of cycles of the full-power carrier for one second, 60,000 will result. Also, if you count the frequency of the reduced-amplitude carrier, 60,000 will result. However, since the carrier power decreases every second, and the duration of the decrease varies in length to impart binary

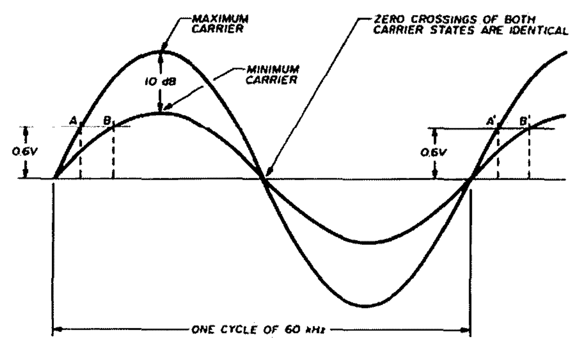
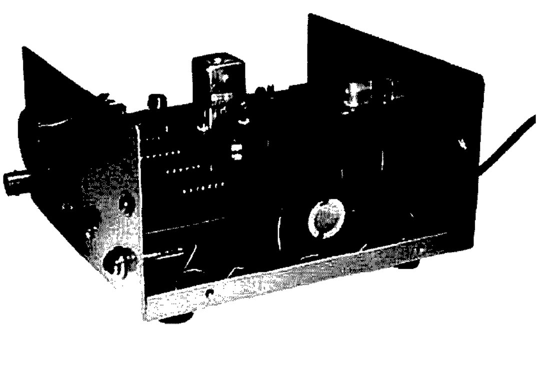


fig. 3. Apparent frequency shift of WWVB when using a digital logic circuit as a 60-kHz counter. The time for 1 Hz, as detected at the 0.6-V level, can be A to A', B to B', A to B', or B to A' depending on the carrier amplitude and the duration of its minimum level.

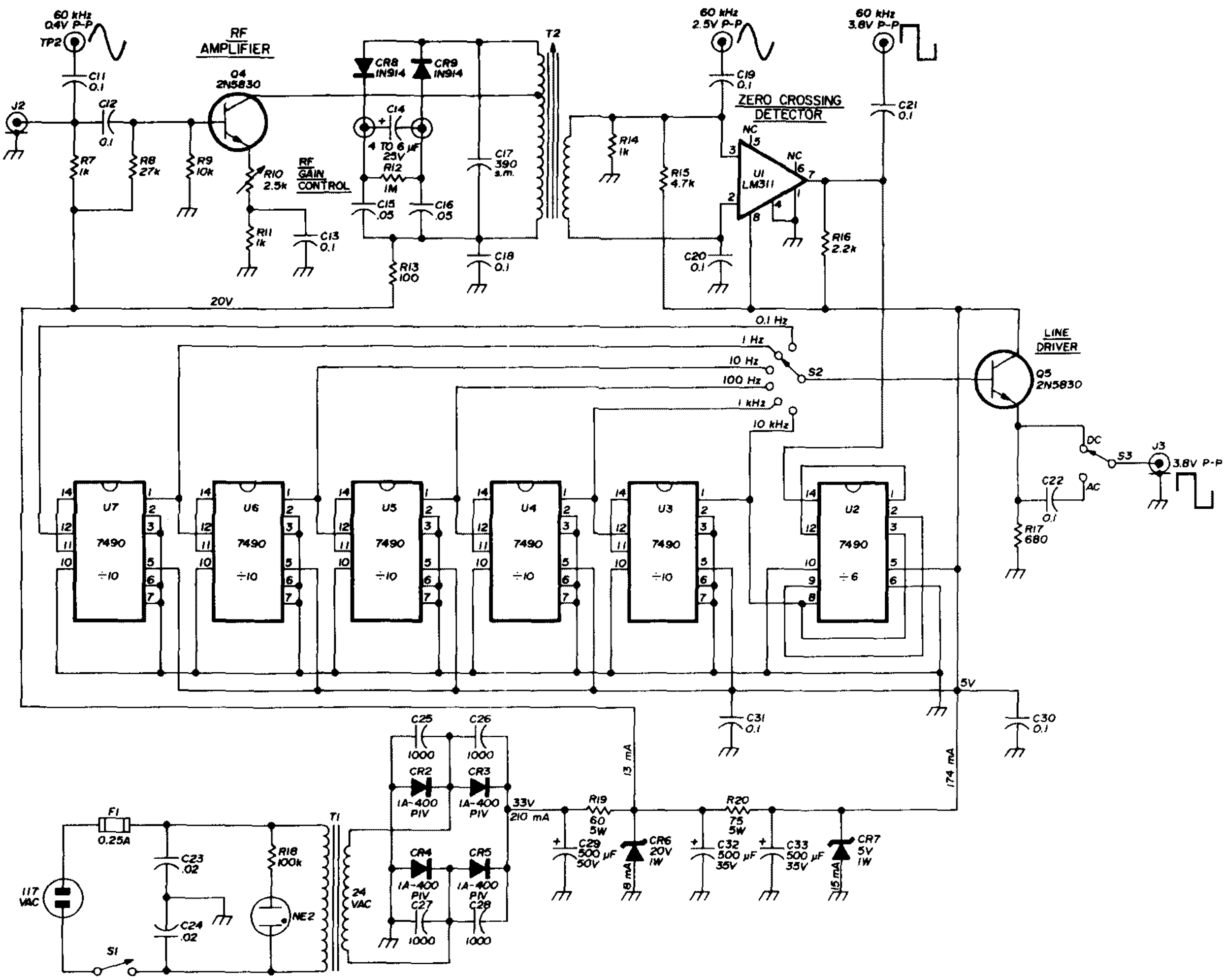
View of the WWVB processor showing part of the divider chain and power-supply wiring.



shown in fig. 3, 0.6 volt occurs at a different point in time depending on carrier power. Since the duration of the shift in carrier power is quite variable, the resulting frequency in the scheme also varied.

Simple and complex limiters were tried in an effort to eliminate the 10-dB carrier-level shift within the receiver. A Bishop limiter was both economical and very effective.[2] A scope observation of the limiter output showed no noticeable change in level, but the counted output showed a -138 to +12 Hz error. There had to be another way.

Further observation of the signal revealed an interesting fact: the zero crossings of the sine waves were always coincident, no matter what the amplitude. So why not count the zero crossings? A search for a suitable zero-crossing detector resulted in the trial of an operational



C14	4 to 6 µF, 25 V tantalum (low leakage)
T1	24 volt, 500 mA power transformer
T2	40 kHz remote control cup-core rf transformer, slug tuned. Total primary inductance, 7.5 to 46 mH, tapped up 0.5 mH; secondary, 20 µH (found in many TV remote controls)

fig. 4. Schematic of rf amplifier, limiter, zero crossing detector, divider chain and power supply.

amplifier. It was a fair detector but its output was not a usable square wave, because op amps are too slow when used in this type of service. An LM311 comparator (fig. 4) was tried with great results; the output was a beautiful square wave. Removal of the limiter further proved that the LM311 was a true zero-crossing detector. (The limiter was retained in the final circuit, however, to correct for possible fading and lightning static.) You'll need a scope with good low- and high-frequency response to observe the square waves within the processor.

Now that we have an accurate and usable 60-kHz square wave, it must be divided down for use as a gate for a counter. A 7490 counter IC connected in a divide-by-six configuration provides a 10-kHz square wave. Five more type 7490 ICs connected as divide-by-ten circuits provide 1-kHz, 100-Hz, 10-Hz, 1-Hz and 0.1-Hz outputs. If your counter has a MacLeish count gate and control logic, you can eliminate one divide-by-ten circuit or switch it out with S2.[3]

The last division by ten is done within your counter by a very necessary multiple-function stage that can't be eliminated. The elimination of one 7490 stage will decrease the 5-volt circuit current by 27 mA, and to maintain 5-volt zener current you'll have to increase R20 to 93 ohms at 5 watts and R19 to 71 ohms at 5 watts (**fig. 4**). You may decide, however, to retain the 7490 stage if you want to use the processor for tasks other than gating a counter; after all, 7490s no longer cost $16.00 each! If your counter uses RTL logic in its divider chain (MC790 etc.), it will be necessary to use an interface device. An npn silicon transistor connected as in **fig. 6** will act as a buffer between the 5- and 3.6-volt systems.

A phone jack with an spst switch attached must be

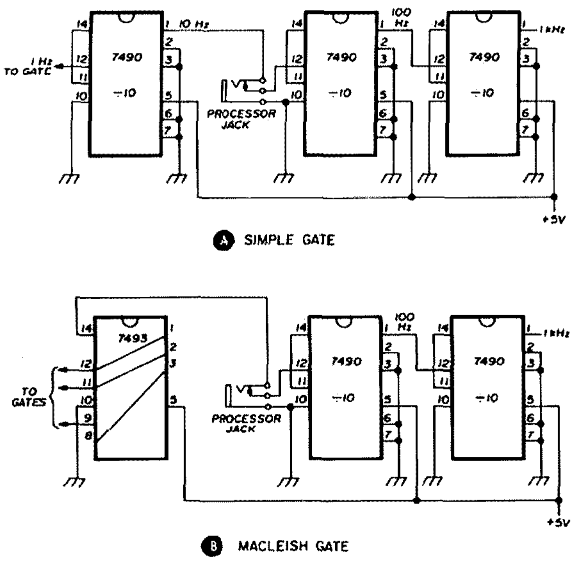
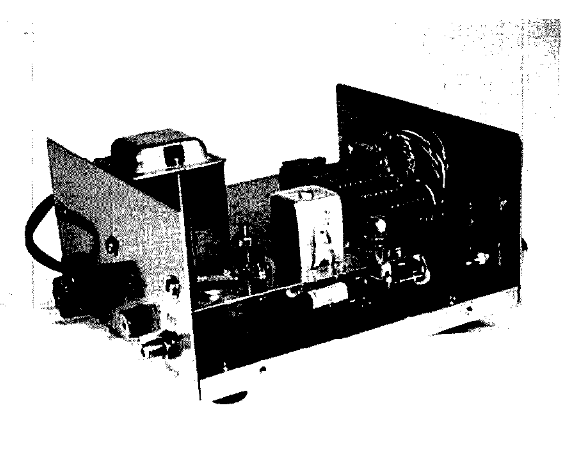


fig. 5. Modification of the two most popular counter crystal oscillator divider chain circuits to accommodate a processor jack.

mounted at the rear of your counter. Connect it as shown in figs. 5 or 6. Those who have built counters will have no qualms about drilling holes, and those with manufactured counters should have no worries either if their work is skillfully done; it's the botched-up appearance of electronic equipment that hurts its resale value!

When the plug from the processor is inserted into the counter, the crystal-oscillator divider chain is interrupted and the processed WWVB signal gates the counter. If the processor output is set at one second, and the counter input is connected to the 60-kHz square wave at test point 4, 60,000 Hz will appear on the counter readout. This is a great test for the health of your counter's gating circuit. Now pull out the plug, allowing your counter to run on its own time base. The readout probably will not be exactly 60,000 Hz at that time.

Left side of WWVB processor showing T2 and LM311 comparator.



Many loop designs are available.[4,5] I chose the one most difficult to build: the double copper shielded type that has immunity to capacitively coupled noise. However, a large loop wound on a simple wooden frame, popular with the broadcast radios of the 1920s, would be fine. A long ferrite loopstick was also tried; it worked but required more gain in the preamp.

Construction details are shown in fig. 1. Obtain two pieces of soft copper tubing at a plumbing supply store. One piece should be ½ inch (12.7mm) ID by 171 inches (4.34m) long. The other, 1 inch (2.5cm) ID by 170 inches (4.3m) long. Before bending into a circle with a 27-inch (70cm) radius, the smaller-diameter tube must be inserted into the larger tube. The small tubing must touch the outer tube at only one spot: the entrance to the preamp box. To prevent a short between the tubes anywhere within the circle, the inner tube is suspended within the outer by liberal use of short lengths of automobile heater hose.

Pieces of hose 2 inches (5cm) long are secured at both

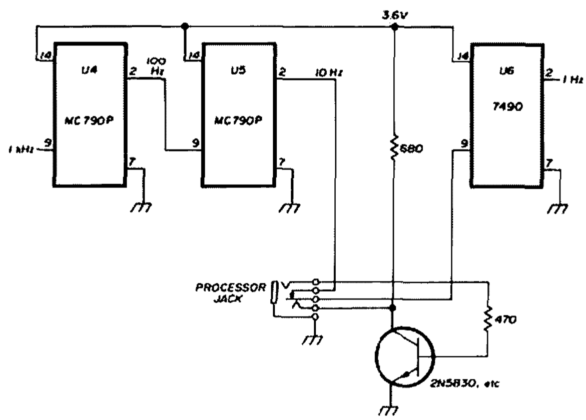


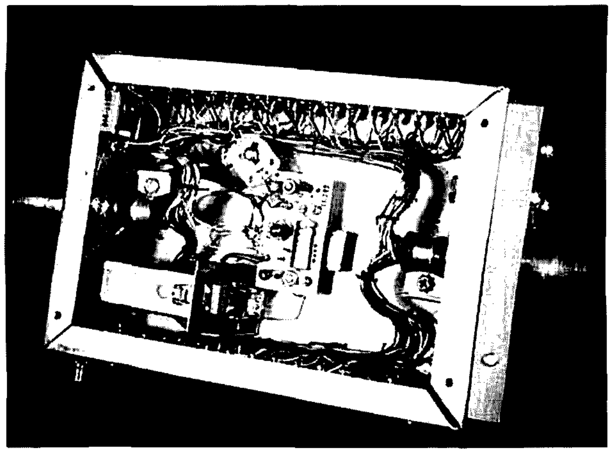
fig. 6. Interface circuit for use between WWVB processor and digital counters having RTL logic.

ends with masking tape to prevent migration from their 12-inch (30cm) intervals during insertion into the larger tube. After carefully forming into a 54-inch (1.4m) diameter circle, you will have to position the inner tube with a mallet. Easy does it! One end should protrude 3/4 inch (17mm), the other 1/4 inch (6mm).

The end with the shortest protrusion is inserted 5/8 inch (15.8mm) into a 1-1/8-inch (29mm) hole punched into one end of the preamp box (see fig. 7). The other end is inserted into a 1-3/8-inch (35mm) hole at the opposite end of the preamp box. This end is insulated by two blocks of fiberglass, formica, or bakelite. The other end is sweat-soldered to the zinc-plated steel box. For this you'll need a torch or a very large soldering iron because the copper is a great heatsink. The inner and outer tubes are connected together at this point with a copper washer soldered in place. Install the insulators after everything cools down. Two large terminal strips with a minimum of 52 lugs total are bolted to the sides of the preamp box.

A 190-inch (4.8m) length of 50-conductor telephone cable was used as the loop inductance. This cable, which is used to connect phones in business offices, seems to be available in short lengths as scrap everywhere. Forty-four of the conductors were connected in series using the terminal lugs within the preamp box. The remaining six conductors were connected in series and used as a low-impedance tertiary winding to couple the high-impedance loop to the low-impedance preamp input. Use care as you connect the cable in series. Use an ohmmeter and observe the cable color coding. One shorted turn can render the loop completely useless.

Forty-four turns of wire resulted in too much inductance, as the coil was easily tuned down as far as 30 kHz. The winding was split in half and both halves paralleled, which resulted in 22 turns of 2.275 mH inductance with a dc resistance of 4.4 ohms. The stray capacitance and a 2500 pF trimmer tuned it to 60 kHz.



Underside of preamp. Terminal lugs are used to series connect a 50-conductor telephone cable, which is used as the loop inductance.

final assembly

The amplifier components, circuit board, test jacks, coax connector and loop trimmer may now be mounted. The finished loop will stand alone, supported by the preamp box. Two wire loops are soldered to the upper part of the loop so that cords can be attached to prevent it from falling over if accidently bumped. The processor and the power supply for the entire system are mounted in a 5-1/4 by 3 by 5-7/8-inch (13.3 by 7.6 by 15cm) equipment cabinet.

The hand-wired circuit layout isn't critical. Three convenient test-point jacks make testing and tuneup easy. The limiter time delay capacitor is mounted on plugs so it can be disconnected during alignment. The 1000-pF disc capacitors, C25, C26, C27, C28 eliminate diode radiation into local vhf receivers. To be 100% effective, these capacitors must be soldered to the diodes with as short a lead length as possible.

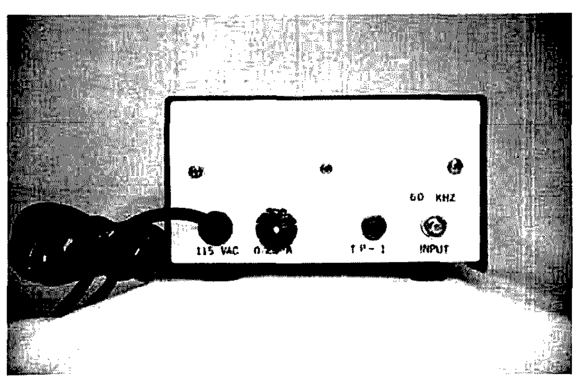
tuneup

After a thorough wiring check connect the preamp and processor with a length of coaxial cable. Any small-diameter coax can be used. This cable also supplies the 10 volts dc for the preamp. Power supply regulation is important. The 10-dB carrier swing and logic dividers tend to modulate the dc supply and may cause a problem in another stage. So be sure that the zener diodes are truly clamped at 5 and 20 volts and that their power dissipation is not approached or exceeded. This can come about because of the wide range in power transformer outputs available to builders.

Signal generators for 60 kHz aren't readily available, so a portable television set may be used as a temporary signal source. Place the television set, which is receiving a clean signal from any local channel, near the 60-kHz loop. Attach an oscilloscope to test point 1 and set its time base to the range that includes 10 to 30 kHz. Tune the loop trimmer and the slug in L3 until a sine wave trace appears. If everything is working well, you are now tuned to about 63 kHz, the fourth harmonic of the television horizontal scanning frequency. The third and fifth harmonics must be avoided. It is sometimes possible to tune homemade loops to 46.65 or 78.75 kHz by accident. If the television set is too close the system may overload as you tune to resonance. It's wise to keep moving the television set farther away as the scope trace increases. Rotating the loop to a partial null is also helpful.

Now turn the television set off. If the loop is oriented for maximum signal from Boulder, Colorado, you should see a weak sine wave with its characteristic one-second dip in amplitude on the screen. Don't forget you're tuned about 3 kHz too high in frequency. Now retune the loop and L3 for maximum using WWVB as a signal source. If the sine wave does not have the one-second dip, one or both of two things are happening. First, there may be a television operating close by – maybe a neighbor's set. Second, the preamp may be oscillating. If the latter is the case, readjust regeneration capacitor C8 until the signal is reduced to the point where the one-second flutter starts. If the preamp continues to oscillate, reverse the tertiary loop winding. The trace, on a well-focused scope, should be a fine single line writing a 60-kHz sine wave. A fuzzy line wider than normal indicates interference from a local television set.

Rear of processor.



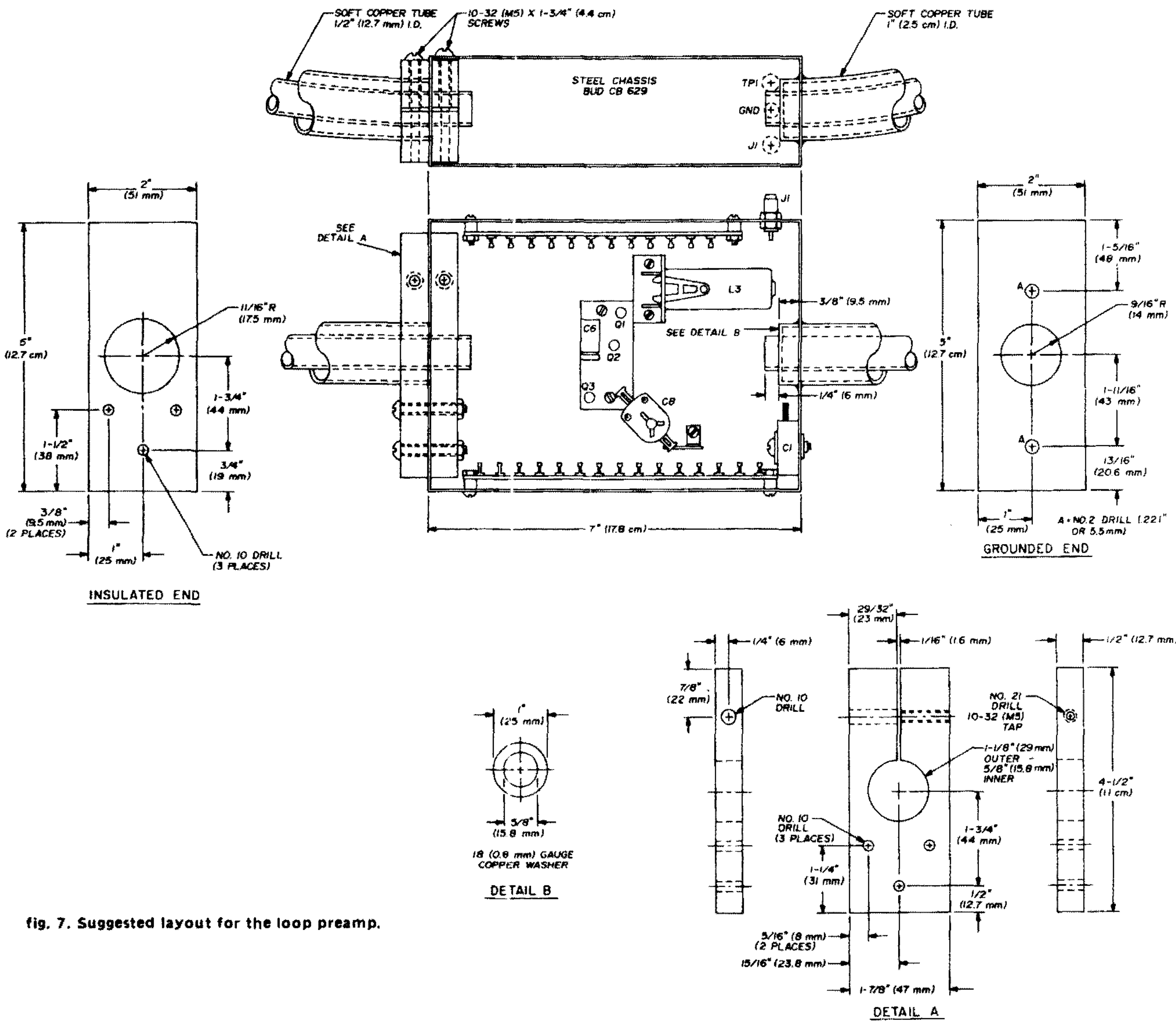


fig. 7. Suggested layout for the loop preamp.

Repositioning the loop so that the interference is at a null while WWVB's signal is maximum is quite successful. I can coexist with the local color television set even though it's only 30 feet (9m) away. System selectivity allows me to live with local television interference. Better than 3-kHz selectivity is not too hard to obtain with three tuned stages at 60 kHz. This will be quite evident when transformer T1 is resonated. Attach the scope to test point 3 and remove time delay capacitor C14. Tune the slug in T1 for maximum. (This is a little difficult because of the one-second flutter.) You only have a second, but it can be done with a little practice. Adjust the gain control at this point for a maximum undistorted sine wave. Reconnect time delay capacitor C14, and within four to six seconds the sine-wave flutter will stop, indicating successful limiter operation.

If your efforts have been successful, you now have a source of accurate timing intervals and frequencies. Although the original intent was to build an accurate time base, the instrument's output may be multiplied for use as a local frequency standard. A 1-MHz output is easily obtained with solid-state multipliers.

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ham radio

high-speed divide-by-n counters

What to expect in available TTL logic for application in two-meter frequency synthesizers

Many articles in the amateur magazines have described frequency synthesizers for use in two-meter fm transceivers. Most generate a harmonic of a crystal-controlled reference frequency by a voltage-tuned oscillator (VTO or VCO) locked to the proper frequency by a programmable frequency divider (divide-by-n counter) and a device that provides a control voltage proportional to the phase difference between the divided-down signal and the reference frequency. The voltage is amplified and used to tune the oscillator to the right spot.

The reference frequency is made equal to the desired channel spacing. For instance, 146.94 MHz is the 4898th harmonic of 30 kHz. If we had a divide-by-4898 circuit and an accurate 30-kHz reference frequency, no great problem would exist in making a stable exciter for an fm transmitter. If we had a switch to make the reference-frequency harmonic anything we wanted between 4867 and 4933, we could do business on any of those fm channels. One difficulty is that such counters (those made from low-priced parts) don't work at very high frequencies. This article describes a counter that goes about as high as possible – at least for fifteen dollars' worth of ICs. It also explains why you can't go higher.

presettable counter IC specifications

In the manufacturers' applications notes I've seen, the highest frequency quoted was 25 MHz for a device with a specified test at 8 MHz. Other literature quotes 22 and 13.5 MHz for input frequencies to the programmable counter portion. However, the Hewlett-Packard HP-8660A synthesized signal generator uses several preset counters operating in the 20- to 30-MHz region. I made a partial copy of the instruction-book circuit. The HP-8660A uses 8290s and 74H logic; I used SN74196s (which were listed in the surplus ads) and Schottky-clamped 74S logic, which I had on hand. It turned out that both were faster than those used in the original.

circuit description

The logic diagram of my circuit is shown in fig. 1. For a three-stage version merely leave off the last decade; maximum frequency remains the same. Details of the first two stages are shown in fig. 1. The counters are preset to, say, 5555. Following the load pulse, which is negative-going on a line that's normally high, the input frequency is counted until the count gets to 9997 (the terminal count). The next clock pulse triggers the auxiliary flip-flop, again enabling the load pulse (two cycles long), getting us back to zero. The total count is 10,000 minus 5555, or 4445 (10-kHz output for 44.5-MHz input).

First I tried the parts barefoot (partially wired in) to make sure they were as advertised. The SN74S112 went

By Hank Cross, W1OOP, 111 Birds Hill Avenue, Needham, Massachusetts 02192

up to 160 MHz in a simple divide-by-two setup. The SN74196s operated as decade counters (the data input pins 3, 4, 10 and 11 grounded) went up to 70 MHz – no problem.

To find out if the unit is working correctly requires enough precision to spot one missing pulse in 10,000. I used either a digital synthesizer or a signal generator and counter for the input signal (−1 to plus 13 dBm sine wave worked okay) and a counter on the output. And of course a calculator is needed. (Lots of digits blinking around.) I found that the divide-by-4445 unit first became a divide-by-4447 then a divide-by-4450 counter as the input frequency was increased until the first-bit decoding wasn't quite fast enough. The 3- and 4-decade versions had about the same maximum count frequency, so I wouldn't expect trouble from adding a fifth decade. One SN74196/8290 (so marked) lost counts when used in the first stage above 28 MHz, but it worked okay in the third. The TI SN74196s went to 56,451.5 with the right numbers. I borrowed some Signetics 82S90s; with these in the first three stages the maximum frequency was 60.89 MHz but was 56 MHz with only one (in the first stage).

There will be some problems at other values of preset, but so far I haven't investigated them, as I haven't figured out how to make 10,000 tests quickly. When the VCO loop is complete and the switch is wired in I'll go looking for trouble. Either the counter reads the same number as the switch or it doesn't.

how to go fast

To make a presettable counter (either up or down counter) work at high frequency we must do two things:

1. Recognize the terminal count and set up the flip-flop before the next cycle of the input frequency.

2. Make a load pulse long enough to ensure that all counters are preset to the right number and the transients have died out. For the SN74196, the book specifies 20 ns; mine was 30 ns at 55 MHz.

To recognize the terminal count, 9997, we look for a 7 in the first decade (0111) and nines (1001) in the rest. Both 7 and 15 have the three-ones pattern in the first three binaries; but an SN74196 does not count to 15, and in any case it would get to 7 first because it's an up counter. The same is true for a 9; two ones are enough to decode a 9. The pattern fills in from the right: the last decade hits 9, then the next, and the next. Four counts later Q_C of the first decade is 1; on the sixth count Q_B is a 1; and one count later the Q_A terminal goes high, about 10 ns later than the input clock's down-going transition. (Both the SN74196 and the SN74S112 work on the negative-going edge of the clock pulse – here it's pretty much a sine wave.)

The best choice for fast action is an SN74S11 positive-AND gate, which gives positive output for all-positive inputs. The pulse, delayed another 4.5 ns by the gate, is applied to the J input of the SN74S112/113/114

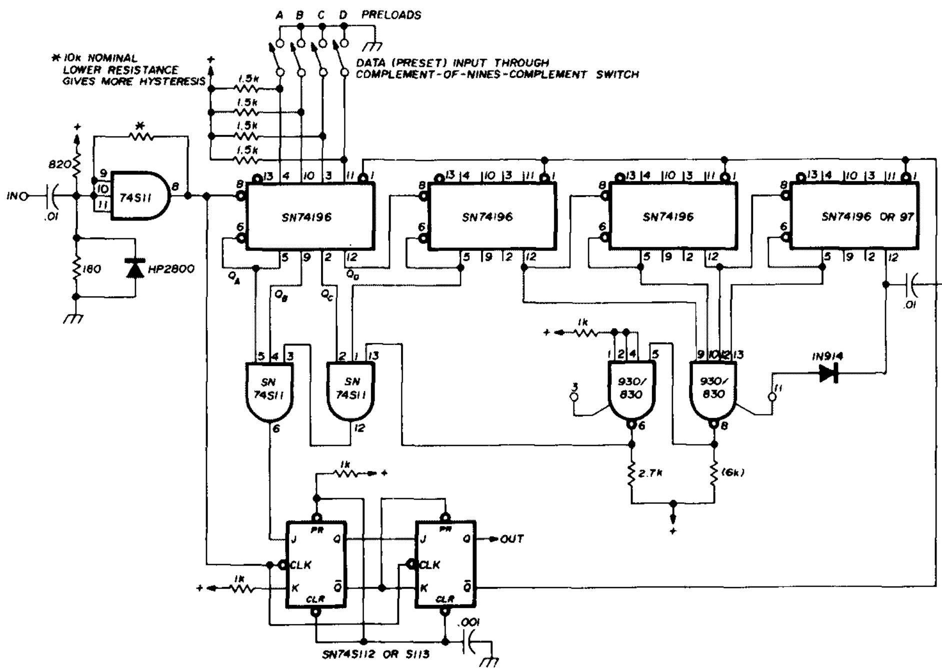
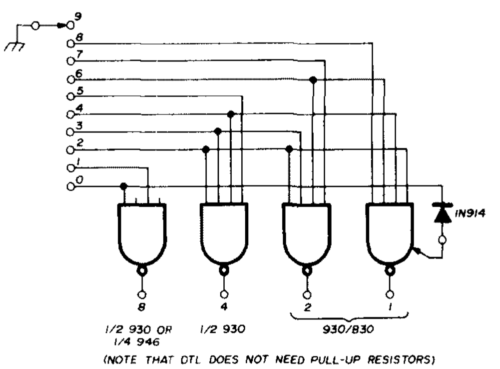


fig. 1. High-speed four-decade presettable counter. "Out" pulse (positive) is 30 ns long at 55 MHz. "Monitor" pulse is 20% duty cycle at about 10 kHz. Input is −1 to +13 dBm sine wave.

dual flip-flop. The specified setup time (J positive before clock transition) is 3 ns, so the total is 17.5 ns, a full cycle at 57 MHz. (The numbers seem to check out.)

Compared with the Q_A output we have an extra 17 ns at Q_B and 52 ns at Q_C, to say nothing of 122 ns for the Q_A of the second decade; so there is only one critical time delay here. If we had used 8 instead of 7 there would have been more problems. A 930 DTL gate does very well for the other decades (see **fig. 1**).

If the gates enable the J input in good time, the SN74S112 toggles on clock number 9998. Q goes up and



DIVIDE BY DIGIT	PRESET (BCD)	1 2 4 8
0	9	1 0 0 1
1	8	0 0 0 1
2	7	1 1 1 0
3	6	0 1 1 0
4	5	1 0 1 0
5	4	0 0 1 0
6	3	1 1 0 0
7	2	0 1 0 0
8	1	1 0 0 0
9	0	0 0 0 0

fig. 2. Nines-complement encoder (one decade; NAND gates used as NORs). The first decade should be tens complement (preset 6 for 4, not 5), which is easily done by a wiring change at left.

$\bar{Q}$ down, working the set (PR) input of the second section, so *its* $\bar{Q}$ goes down. This output is the load input of all four decades. On count 9999 the first section returns to normal. On count 0000 the second section returns to normal, ending the load pulse. At 57 MHz this is two periods (35 ns) less 5 (the extra time for the direct set to work), or 30 ns. The load pulse extends into the first count period by 5 ns, but this doesn't seem to make any difference.

some device advice

My experience with available devices used in high-speed counters is presented below for those interested in what to expect. The TI SN74196, Motorola MC4016, and TI SN74192 are examined.

Up counters are easier to decode and are more difficult to preset than down counters such as the MC4016 and SN74192. The SN74196 presents some problems. Preset coding is by nines-complement (9 minus the desired number). The best switch to use is called "complement-of-nines-complement," but these seem hard to come by. For a few digits, an encoder from 10-point to nines complement (or any other code) can be made with four gates per decade using two 930s, as in **fig. 2.** A BCD nines-complement converter can also be made using gates, inverters, and an exclusive-OR per decade.

The SN74196 uses external terminal-count decoding and the counters are nonsynchronous. On the other hand, it operates as a preset counter beyond 50 MHz.

The MC4016, designated as a divide-by-n down counter, has other problems. It costs as much as an 82S90. Its Q_A delay, although not specified, is about 20 ns. Its Q_B delay is high, and its built-in zeros detector is not usable in the first decade. Its maximum frequency is probably less than 35 MHz with all possible circuit tricks employed. Early decoding should probably be done on "4." Three flip-flops would be required.

The most popular down counter is the SN74192. Its price is right; a BCD or (preferably) a BCD-complement switch presets it; it has built-in terminal count decoding ("borrow"); and the counters are quasi-synchronous. Time delay at the Q outputs is specified, although delay is longer for a 1-0 transition.

BCD down counting means that either the first bit goes high or zero, or that two bits change together, which involves a 1-0 change thus limiting decoding speed. Minimum decoding delay of the SN74192 is about twice that of the SN74196. So the maximum frequency of the SN74192, based on book data, would probably be no more than 28 MHz; however, I haven't made any circuit measurements in this regard.

construction

I made tests on a circuit wired on a 3677-2 Vector board. This board costs more than the ICs, with its socket. I used Cambion sockets for ease in swapping counters. The Vector board has ¼-inch (6mm) wide strips for B+ and ground. I used several CS13 tantalum-slug bypass capacitors between the two buses at strategic points. Wiring was no. 22 (0.6mm) solid hookup wire. I found no layout or grounding problems. All unused inputs were connected to V_{CC} through a Ik resistor. You'll never find out what's going on without a high-speed dual-trace scope; however, once it's going, you're all set.

The input circuit in **fig. 1** worked better than several fancier "trigger" circuits. Note that a resistor between the input and output of an SN74S11 provides positive feedback. Without the resistor the circuit may oscillate with no input signal; adding positive dc feedback makes it a trigger with hysteresis. The whole counter draws less than 300 mA at 5 volts.

Something that hits 10-kHz multiples at 48 MHz has obvious application on two meters. That's the next step.

ham radio

superfluous signals

On-the-air tune up is becoming more frequent. These ideas will help you avoid this illegal practice

Amateurs have recently increased the practice of tuning up on the air, which has reached such a level that it's often impossible to work a DX station or participate in a net. Many tune-up signals are weak, but others are far stronger than the desired signal. Such practices are not only annoying, they're also illegal. An excerpt from the International Telecommunications Union regulations (693) states in part *"All stations are forbidden to carry out . . . unnecessary transmissions . . . the transmission of superfluous signals and correspondence . . . (and) the transmission of signals without identification."*

Observation and inquiry show that three factors are involved in the development of on-the-air tune up practices: the characteristics of modern transmitting equipment, lack of knowledge about corrective steps, and the "me first" approach too common today. This article is devoted to the second factor, corrective steps, with some discussion of the reasons why they are necessary.

One of the characteristics of many modern commercially built exciters, transceivers, and linear amplifiers is that they lack dial scales. Some may have a mark at 9, 12, and 3 o'clock, or even at every hour position, but many are blank. As a result, it's almost impossible to preset operating controls with any degree of accuracy. The only way to obtain proper performance is to tune up with *reference to meter readings. Since amateurs* want full output, it's no wonder that on-the-air tuneup has become so common.

The solution to this problem is simple – install useful dials. Here, useful means a dial that can be read to about one degree of arc or so. It certainly means one that's large enough to get your fingers on, is easy to read, and has no backlash. For many controls it also means a vernier knob.

Most users of commercially built equipment have put up with the lack of useful dials from fear of spoiling equipment resale value. This shouldn't be a worry, because it isn't necessary to change appearance, drill holes, or take other steps that will reduce resale value. Also no great amount of work is necessary. On most transmitters or amplifiers only three controls are involved: typically drive, tune and load. A couple of hours of work should take care of most cases.

The simplest approach is to remove the old knobs and replace them with dials having calibration marks close enough to be useful, typically 0-100 over an arc of 180 degrees. Many styles are available in junk boxes, surplus outlets, and stores. However, it's difficult to find small dials with the necessary calibrations. If you're fortunate enough to obtain such dials, put the removed knobs in a cloth bag and tie it inside the equipment where it won't get lost. This saves no end of trouble when it's time to sell or trade.

Another simple approach is to install a calibrated dial plate under the control mounting nut. These plates were

By Robert P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

common for many years but are scarce now. They can be made using dry-transfer kits similar to those for lettering, or the center portion of a piece of polar graph paper, suitably lettered, can be used to give calibrations each 2 degrees. Dial plates can also be hand drawn. Best appearance is obtained by making the drawing several times full scale then reducing it photographically.

Both these methods have one disadvantage: they don't have vernier action. Small vernier dials are usually available from Radio Shack, Olson or Lafayette in 1½, 2, and 3 inch (3.8, 5.1, and 7.6cm) sizes. These dials can be mounted on a false panel section using countersunk

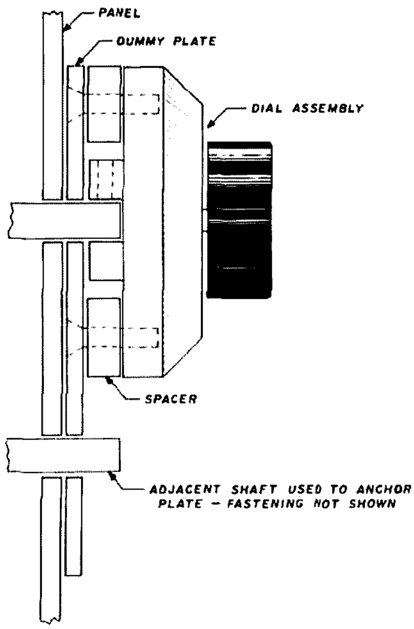


fig. 1. No-holes dial mount using a dummy plate.

screws. The assembly can be held in place by the control nut or by an adjacent nut as shown in fig. 1; by bending around the edge of the panel; or by a drilled hole hidden by the original knob. *QST*'s "Hints and Kinks" contain other mounting ideas.

Another method, which involves considerable work but which offers the opportunity to obtain a "custom station," is to make a complete dummy front panel. Custom features such as combinations of equipment, special lettering, engraved calls, and so on can be had by this method. It's a good method for organizing the clutter of auxiliary controls that most stations accumulate. Whatever the method used, an additional item needed is a setting log or card giving the proper setting of each control for each preferred frequency or for band segments on the low frequencies and bands on the high frequencies. Plastic envelopes are convenient for these logs.

antenna tuner

The ability to return to a known dial setting is a big

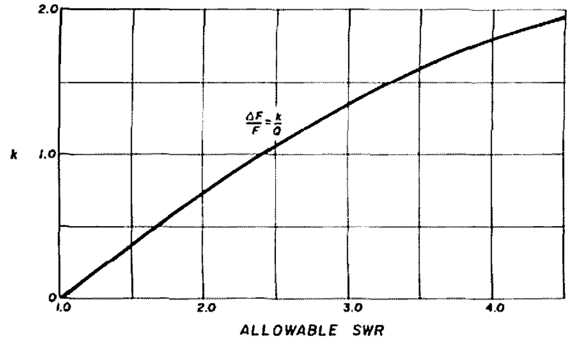


fig. 2. Bandwidth corrections for allowable swr.

help but doesn't fully compensate for another characteristic of modern transmitters and amplifiers, the fact that they're designed to work into a 50-ohm load with an swr no greater than 2:1. An easy way to see how restrictive this can be is to consider the effect of antenna Q. The Q will be around 70 for a typical close-spaced Yagi on 7 MHz. The response will be nominally within 3 dB for a change of ±50 kHz. However, the Q response curve must be multiplied by a factor, *k*, which takes into account the allowable swr. The way this factor varies with swr is shown in fig. 2. The maximum range of transmitter adjustment is used up with a frequency change of only ±37 kHz. Even with low-Q antennas such as a wide-spaced Yagi with a Q of around 7, where the transmitter adjustment range covers 350 kHz or so, there will be measurable difference in transmitter performance with a frequency shift of 30 kHz. No wonder on-the-air tuneup has become so common!

If we look at the reason for this, we find that the

Dummy antenna load fashioned from a paint can. Ten 510-ohm 2-watt resistors in parallel, immersed in mineral oil, dissipate 470 watts for 5 seconds — adequate for tune up.

culprit is the reactance caused by operating away from resonance. For example, the resistance and reactance values for a typical dipole are shown in **table 1**. For a one-per cent change in frequency, (70 kHz at 7 MHz), the radiation resistance changes by about 6 ohms, or less than ten per cent. The reactance change is over six times as large numerically.

The solution to the problem introduced by this reactance change is simple — use an antenna tuner. Furthermore, since antenna reactance causes most of the swr change, we can use a single variable element. The other antenna tuner elements can be fixed. The improvement possible by this reactance control method is evident from fig. 3. The upper curve is the swr expected with a typical wire dipole fed with a 50-ohm line. The swr reaches 2:1 for a one-per cent change in frequency. By simply cancelling the reactance, the swr seen by the transmitter changes to the lower curve. Its minimum swr is now lower, reaching 1.0; and 2:1 is reached only after a 5½ per cent frequency change. Note that the minimum swr occurs at a different frequency as far as the transmitter is concerned. Note also that the swr on the line does not change: it reaches about 7:1, causing increased voltage across the line, which may cause breakdown with small line and high power. The high swr also causes a small increase in line loss, usually entirely negligible. The

table 1. Impedance of a typical dipole antenna.

per cent resonant length	resistance (ohms)	reactance (ohms)
46	47	−95
47	53	−76
48	59	−38
49	65	0
50	71	+38
51	77	+76
52	83	+95

gain in frequency flexibility is far more important than these small problems.

The easiest way I've found to attain this single control action is shown in fig. 4. The pi network is conventional, with the arms continuously adjustable. The difference is in its use. L, C1 and C2 are preset to the band used and only C1 varied as frequency is shifted within the band. The indication for proper adjustment is a zero reading on the reflected power meter, which is an *ARRL Handbook* meter with only the reflected power elements connected. C1 is first preset to the table value, then adjusted as required when transmission begins. For a matched load, proper adjustment occurs with the reactance of each arm equal to the line impedance. This also gives the element values for a switched arm network.

dummy load

Good dials and a single-control matching unit are a big help in solving the tune-up problem, but it's still comforting to know that everything is tuned up on the nose. We can have this capability without radiating by adding another element — a dummy antenna automatically inserted into the transmitter output when the

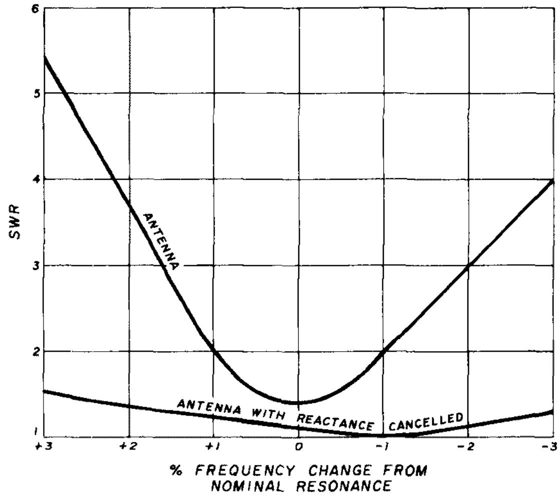


fig. 3. Typical swr for dipole antenna with and witnout corrective steps.

transmitter *tune* control is activated. A circuit for this is shown in **fig. 5**. The diodes prevent interaction between the external relay and the internal circuits. The switch allows the test signal to be radiated, primarily to determine initial matching network settings.

The dummy antenna can be a commercial or kit unit, but the large units rated at 1 kW are not needed in this service provided that tuneup is held to the few seconds needed to touch up preset controls. A dummy antenna isn't hard to build; the basic components are a handful of resistors. The following discussion is based on use of Allen-Bradley resistors. Units of other manufacture may be satisfactory, but the rating values should be secured from the manufacturer.

A 2-watt A-B resistor of the twenty-per cent series is rated for a continuous load of 2 watts for 100,000 hours when the resistor is mounted with 1-inch (2.5cm) leads and has a body temperature of 212°F (100°C): this occurs when the ambient is 122°F (50°C). The life rating increases by a factor of ten for a 122°F (50°C) reduction in body temperature. The allowable load increases by 40 per cent for a 10:1 reduction in life. For short-term loads the resistors, for the same mounting, ambient and life, are rated at 44 watt-seconds; i.e., 44 watts for one second.

Suppose one of these resistors is used as a dummy antenna for a five-second tune-up checks. Over a ten-year period, if used ten times a day, the load would be applied for about 50 hours total. Since the required life

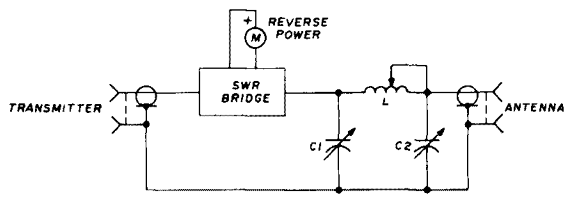


fig. 4. Antenna tuner for reactance compensation.

under load is so short, the load can be increased accordingly. Also, the resistor body temperature can be decreased by mounting the resistor on heavy fins that touch the ends of the body and by immersing the fins and resistors in cooling oil. The body temperature will then be essentially ambient, certainly no greater than 122°F (50°C).

As a result of these steps, the power input can be increased by 1.4 times for the reduction in temperature, and 1.4^4 times for the acceptable reduction in life, or by 5.35 times. The rating is now 10.6 watts continuous, or 47 watts for five seconds. Ten 510-ohm resistors in parallel will dissipate 470 watts for five seconds; easily full output for a 600-watt-input transmitter of the type that runs at full output on tune, and ample for a 1 kW unit that tunes at 50 per cent of full output. Forty resistors will handle a maximum legal input transmitter.

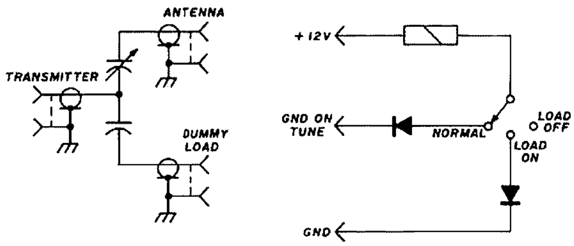


fig. 5. Circuit for dummy-load control during tune up.

A dummy load designed in this fashion is shown in the photo. The container is a one-quart paint can. Fins are copper. The 2-watt resistors are separated by three body diameters in each direction, and are staggered for good oil circulation. The cooling oil is light mineral oil, but transformer oil or hydraulic fluid are usable; mineral oil is odor and stain free. The unit shown is eight years old and has been used for tests up to 30-seconds duration with a 500-watt-input transmitter, repeated until the can became quite warm. Actually the rating method above seems to be quite conservative, since the change in resistance in this period has been much less than the 10-20 per cent expected.

the operator

With these three aids in place it's now possible to get on the air, correctly tuned up, without causing the slightest interference. The steps are:

1. Set the exciter, final amplifier and tuning dials to the logged settings for the band.

2. Activate the tune control, placing the transmitter on the dummy antenna, and touch up the tuning if readings aren't normal.

3. Start transmitting, observing the antenna tuner reflected-power meter; adjust the antenna tuner if needed.

These steps will result in the maximum possible signal and will also make life more pleasant for others on the channel.

ham radio

vhf/uhf techniques

Joe Reisert, W1JAA

what's wrong with amateur vhf/uhf receivers — and what you can do to improve them

How many times have you heard the expression, "If you can't hear them, you can't work them?" Well, this old saw pretty well sums up one of the biggest problems faced by amateur vhf and uhf operators. The culprit usually turns out to be the receiver or converter.

A complete treatment of the subject of vhf and uhf converters would require a lengthy article which is beyond the scope of this column. Therefore, I will limit my remarks to a general nature and, wherever possible, use references for more detailed information. Furthermore, I will only direct my remarks toward vhf/uhf converters; I'll leave the i-f or tunable receiver up to the user. Hopefully the remarks in this column will stimulate some interest in upgrading your converters, and will make your vhf/uhf operating more enjoyable.

general converter problems

The primary problems with most vhf/uhf converters are high noise figure, spurious responses, gain compression and "intermods" or IMD (intermodulation distortion), poor stability, and burnout. I'll briefly describe each of these problems, suggest some cures, and finally, recommend some tests to measure converter performance.

High noise figure is a symptom of almost every vhf or uhf converter I have ever measured. Principal causes (disregarding defective components for the moment) are high noise figure in the preamplifier, low preamplifier gain, high conversion loss in the mixer, and high i-f noise figure. Combinations of the above problems are also prevalent — the overall effect is poor sensitivity.

Spurious responses are also quite common in vhf and uhf converters. Unfortunately, instead of eliminating the spurious problem, most operators learn to live with it. Typical causes are insufficient filtering in the rf and local oscillator circuits, poor choice of intermediate frequency, and instabilities in either the preamplifier, oscillator, or frequency-multiplier stages. The overall result is the appearance of undesired signals in the frequency range of interest.

Intermodulation distortion and gain compression are usually caused by insufficient dynamic range at some point in the converter or i-f. The most likely culprits are poor linearity or excessive rf gain, or both. The requirements for low noise figure enhance the problem since rf gain and lower device operating currents are necessary. The usual result is the appearance of undesired signals which cannot be eliminated unless attenuation is placed ahead of the converter.

Poor stability, like spurious responses, is a problem many vhf/uhf operators learn to live with. It almost goes without saying that the i-f following the converter must be stable. The usual causes of instability are the oscillator circuit (and its components), the power supply, and the physical surroundings of the oscillator. Poor stability will cause the received signal to drift or rapidly change frequency.

Burnout is a perennial problem which has increased with the widespread use of solid-state devices. It can be caused by electrical discharge (such as lightning), high rf signal levels, or improper power supplies -- anything that overstresses the devices used in the converter. The typical result is a dead converter without any prior warning.

Before discussing any specific circuit recommendations, I would like to give my usual pitch for the *modular* approach to vhf/uhf converter design where virtually every stage or function in the converter occupies a separate chassis or enclosure. There are many advantages to this system. First of all, modular converters are easy to work on because you can work on one circuit at a time — none of the other circuits is affected. Why build a new local oscillator chain every time you build a new converter or preamplifier? Furthermore, optimum performance of each module can be attained on your bench before you test the circuit in your converter. Shielding between stages is virtually assured with separate modules so crosstalk is vastly improved. Finally, if any module fails, it can usually be rapidly bypassed so operation will not be completely curtailed or severely degraded.

The primary disadvantages to the modular approach to vhf/uhf converter design are size, complexity and cost. However, if optimum performance is your goal, the advantages far outweigh these shortcomings. Just think how easy it will be to try out a new preamplifier or mixer if you use the modular system. All it takes is a simple substitution of modules, a one- to two-minute operation. If the change is not productive, the converter can be restored to its original configuration in seconds. You no longer have to build a complete new converter just to evaluate a new circuit.

A block diagram of a typical vhf/uhf converter is shown in fig. 1. If you use the modular approach, up to ten individual modules units could be used, but usually some units such as the crystal oscillator and frequency multipliers can be combined in one module.

recommendations

The overall noise figure of a well designed vhf con-

verter is usually determined by the preamplifier. The gain of the preamplifier should typically be 10 to 12 dB, minimum, but not greater than 18 to 20 dB. Most vhf/uhf preamplifiers with more than 18 to 20 dB gain are only conditionally stable (many oscillate if terminated in an impedance other than the design value – usually 50 ohms). As a rule of thumb, for high dynamic range the stage gain should be about 6 dB greater than the noise figure of the system following that stage. For low-noise converters stage gain should be 10 dB greater than the noise figure of the following system.

If the mixer has a 5 dB noise figure, for example, 15 dB of preamplifier gain would be desirable for minimum noise figure; 11 dB gain would be sufficient for a high dynamic range converter. Notes on calculating gain and noise figure are contained in the appendix of reference 1. Reference 1 also discusses the benefits of low noise preamplifiers and illustrates recommended circuits and devices.

Although it has been stated many times in the past, it should be reiterated that maximum gain and lowest noise figure are rarely coincidental in preamplifiers. In addition, the input filter to a converter should be a low-loss, single resonator type such as a quarter-wavelength coaxial cavity[1] because any loss ahead of the converter will be added directly to its overall noise figure. Multiple-resonator input filters should be avoided because the input mismatch to a typical low noise figure preamplifier may cause the input tuning to change, thereby increasing the noise figure.

Quarter-wavelength input filters should be tuned for minimum vswr into a well matched 50-ohm load. Then they should be connected to the preamplifier with a *short* connection (such as a coaxial adapter) and *never be retuned*. Retuning a filter after it has been connected to a preamplifier frequently results in increased insertion loss unless precision noise figure measurements are used.

Other sources of high noise figure are high mixer conversion loss and poor i-f noise figure. A suitable low loss mixer will be discussed later in this column. The typical i-f used by most vhf/uhf operators is a commercial 28- to 30-MHz communications receiver. It is not unusual to find that the noise figure of such a receiver is as high as 10 or 20 dB! Therefore, it is recommended that a low noise figure preamplifier be placed ahead of the receiver. This will also reduce the need for high preamplifier gain, thereby increasing dynamic range. A suitable i-f preamplifier is discussed in reference 2.

As pointed out earlier, spurious responses are quite common in vhf/uhf converters. Careful choice of the i-f is a very important determining factor in this regard. Generally speaking, 28-30 MHz is an optimum choice. Lower intermediate frequencies are more susceptible to i-f breakthrough and increase the complexity of filtering out the image frequency. Reference 3 provides some good guidelines for choosing an i-f. Image rejection of at least 30 dB should be a minimum goal.

Filtering of the rf and local oscillator circuits should be carefully considered since insufficient filtering will enhance spurious responses. Proper filtering of the local oscillator is seldom given adequate attention in most amateur converters, and subjects the mixer to additional undesired frequencies. The result is similar to the computer programmers' expression, "GI-GO" (garbage in – garbage out).

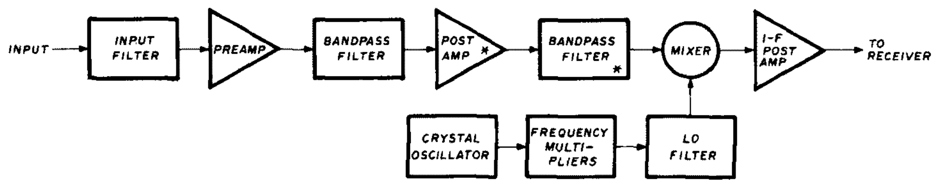


fig. 1. Block diagram of a typical vhf or uhf converter. The blocks marked with an asterisk are not necessary if high dynamic range and moderate noise figure are desired. The frequency multiplier is needed only in those cases where local-oscillator injection is above about 125 MHz (the upper limit for recommended overtone crystals).

Interstage and local-oscillator filters should preferably have multiple resonators (tuned circuits) to enhance selectivity. Losses of 1 to 2 dB are usually tolerable since there is generally adequate gain available to compensate for filter losses. Interdigital filters[4] are preferred for uhf while comb-line and helical resonator types[5,6] are preferable for vhf.

A fundamental oscillator frequency of 90 to 125 MHz is recommended for all converters for 144 MHz and above. Lower frequency oscillators generally cause more spurious frequencies, are more difficult to adequately filter, and can cause unwanted reception of television and fm broadcast signals. A suitable oscillator circuit is shown in **fig. 2** and will be discussed later in this article.

Frequency doublers are preferred for multipliers since they are more efficient (especially the push-push type), more stable, and easier to filter. Frequency triplers, however, are acceptable when the output is below 300 MHz. Typical doubler circuits are shown in **fig. 3**.

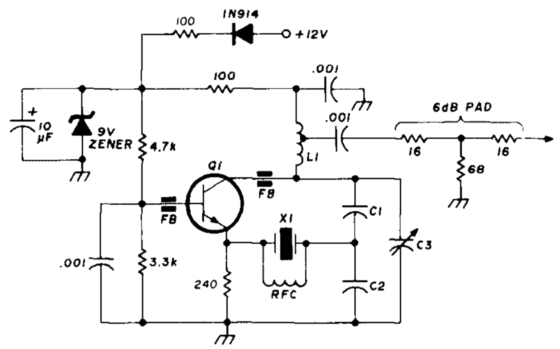
Another cause of spurious outputs in vhf/uhf converters is instability in preamplifiers and local-oscillator/multiplier stages. Care should be taken to adequately decouple and shield all circuits in the local-oscillator chain to prevent this type of instability. The modular approach is a definite benefit in this regard.

intermodulation distortion

Gain compression and intermodulation distortion are particularly prevalent in most of the vhf/uhf converters I've seen. Gain compression is simply the lack of dynamic range with a signal at the input. Intermodula-

tion distortion, which must not be confused with spurious responses, is the result of mixing the harmonics of one or both of the two input signals which in turn generate other frequencies within the desired passband. This is caused by non-linearities, either in or ahead of the mixer, and has been well documented.[7,8]

First of all, you must insure that only the desired signals enter the converter. This can usually be accomplished with narrowband filters at the converter input and throughout the rf paths as discussed previously. If



C1	10 pF mica
C2	20 to 60 pF mica. Use as high value as possible (until circuit just oscillates reliably when C3 is tuned through resonance)
C3	20 pF piston or miniature trimmer
L1	8 turns no. 24 (0.5mm) on Amidon T37-12 toroid core, tapped 3 turns from cold end
Q1	Fairchild 2N5179 recommended but 2N2857, 2N3563, 2N918 or equivalent may be substituted
RFC	0.39 µH. Resonates with crystal holder capacitance (4 to 6 pF typical) for parallel resonance at crystal frequency
Y1	90 to 125 MHz, 5th or 7th overtone, series-resonant, HC-18/U crystal. Cut leads as short as possible (¼" or 6mm maximum)

fig. 2. Recommended 90 to 125 MHz crystal oscillator circuit for vhf/uhf converters. Output is 5 to 15 mW. Crystal should be a high-quality 5th or 7th overtone type. The ferrite bead (FB) prevents undesired oscillations above 500 MHz.

the undesired signals are too strong or too close to the input frequency, additional precautions must be taken.

The mixer is usually the major source of IMD difficulty since it is essentially a non-linear device and sees all input signals that remain after filtering and preamplification. The best solution to mixer-generated intermods is to use a mixer which has good dynamic range. Bipolar transistors are extremely poor in this regard and should be avoided at all cost. Fets, including the dual-gate mosfet types, are good at vhf but are poor performers above 300 MHz.

The double-balanced mixer is recommended since it generates fewer spurious responses than single-ended or single-balanced designs. The fet double-balanced mixer is recommended for operation below 300 MHz because it has good dynamic range and conversion gain.[9] For all around versatility, however, the Schottky-diode double-balanced mixer is hard to beat since inexpensive units (less than $7.00 in single quantities) cover the frequency range from 5 to 500 MHz and present 50 ohms at all ports. The circuit of one popular type is shown in **fig. 4.**

For best intermodulation performance the Schottky double-balanced mixer should have sufficient local-oscillator injection (10 to 20 milliwatts) and all ports should be properly terminated. The recommended procedure is to use a 3 to 6 dB resistive attenuator on the local-oscillator port (with a commensurate increase in local oscillator power), and a well matched i-f preamplifier on the i-f port, preferably with a diplexer. This subject is thoroughly discussed in reference 9. An extension of this technique would be to use one of the high dynamic range double-balanced mixers. They require even higher local oscillator power but the higher mixer outputs may cause the i-f to generate intermodulation products so their use may not necessarily justify the added expense.

The gain problem is a whole, separate subject. I have noticed all too often that far too many preamplifier stages are used ahead of the mixer. Generally speaking, two stages, (one preamplifier and one postamplifier) should be more than sufficient, even for the lowest noise receiver. Gains ahead of the mixer totalling 30 dB or more are definitely undesirable. High gain rarely contributes any worthwhile performance benefits and decreases the dynamic range of the receiver.

The linearity of the postamplifier (the amplifier stage after the mixer) is also very important. As a general rule of thumb, this stage should operate with a collector (or drain) current of four to five times the corresponding current of the preamplifier. This is necessary because the signals present in this stage are typically 10 to 15 dB higher (gain of the preamplifier less loss of the filter) than at the receiver input so the postamplifier is more susceptible to non-linearities. It may be worthwhile to note that for *each dB of extra gain ahead of the mixer the intermodulation distortion increases by 3 dB*! If lowering the gain ahead of the mixer causes i-f problems (poor noise figure, etc.), a postamplifier can be added after the mixer.

stability

Frequency stability can sometimes be a problem. The principal causes of drift are the actual oscillator circuit, its components, and, in particular, the crystal. Only high quality components and overtone crystals should be used. Inexpensive crystals should not be used. Inexpensive crystals frequently generate spurious outputs, noise, and drift. Drive levels to the crystal should be kept to a minimum commensurate with reliable performance. Frequency-pulling components such as trimmer capacitors should be avoided at all costs as they will seriously degrade the stability and spurious performance of even a good high-Q crystal. The crystal should be allowed to operate at its natural series-resonant frequency, and the oscillator power supply should be regulated. A recommended circuit is shown in **fig. 2.**

Needless to say, heat-generating devices such as vacuum tubes should not be located near an oscillator because they will seriously affect stability. Furthermore, all oscillator components should be rigidly mounted to prevent any mechanical instabilities.

Burnout has increased with the use of higher power transmitters and solid-state converters. All circuits, as a minimum, should use an *idiot* diode (a diode in series with the power supply as in figs. 2 and 3). They prevent damage due to polarity reversals and lower crosstalk. Regulated power supplies for solid-state equipment are highly recommended and inexpensive due to the ready availability of three-terminal voltage regulators.[11] A power supply which is used for relays should never be used to power a solid-state circuit because relay coils induce large voltage spikes on the power supply line during switching operations. It shouldn't be necessary to mention it, but never operate a circuit at voltages (or power dissipation) above the manufacturer's maximum specifications.

Other forms of burnout are excessive rf input or transients. The ordinary coaxial relays that are used by vhf/uhf amateurs often have only 30 to 40 dB isolation between the transmit and receive ports. If you're running a high power transmitter (500-600 watts output) this means that up to 0.5 watt of rf power can be impressed across the receiver input.

One simple cure to this problem is to use an additional lower power relay in series with the receiver for additional isolation. Another solution is to use a limiter ahead of the preamplifier. Even a single silicon or hot-carrier diode across the base-to-emitter junction of a bipolar transistor will help (see fig. 5). Germanium or back-to-back diodes across the input line should be avoided since they frequently turn on too early and can also generate spurious responses from out-of-band signals.

Out of band transients and excessive rf can usually be eliminated with a good input filter as discussed previously. It is not uncommon to see *volts* of rf energy on the transmission line from a vhf/uhf antenna which is near a high-frequency transmitter such as a 10-80 meter amateur station.

Lightning is also a real problem. If two relays are used (as described above), they can be connected so the preamplifier input is terminated in a 50-ohm dummy load when the receiver is not in use (see fig. 6). A short-circuit or open return may cause a preamplifier to oscillate. This oscillation could also cause damage to the device and may generate local, self-caused interference.

Another possibility for lightning protection is a highpass filter which is designed with a cutoff frequency which is 10% to 50% below the band of interest (see fig. 7). Since most of the energy in a lightning strike is concentrated below 50 MHz a highpass filter will provide a measure of protection against nearby electrical storms.

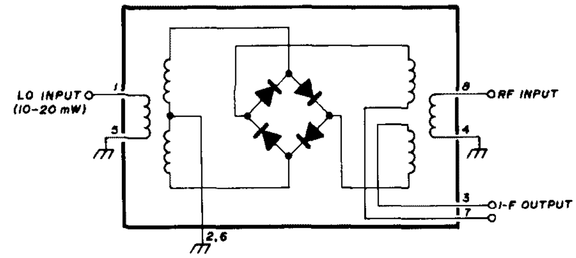
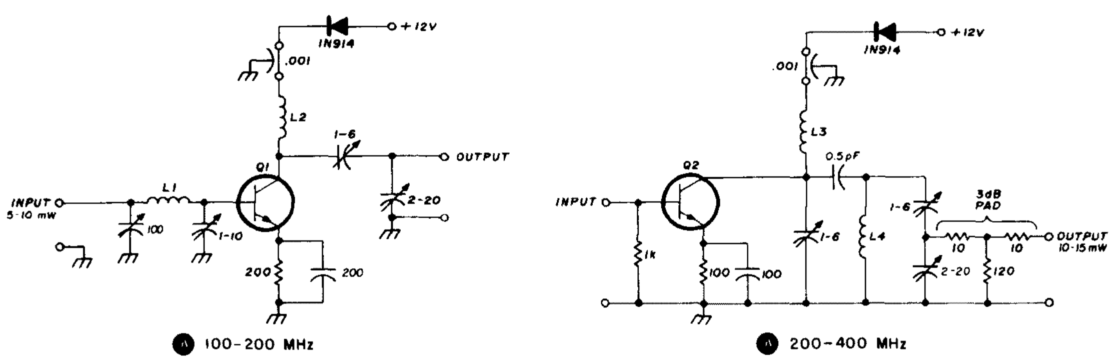


fig. 4. The Anzac MD108 double-balanced mixer is useful over the frequency range from 5 to 500 MHz and is recommended for amateur vhf/uhf converters. This device is priced at $7.00 in small quantities (plus postage) and is available direct from the manufacturer, Anzac Electronics, 39 Green Street, Waltham, Massachusetts 92154.

However, there is virtually no protection against a direct hit!

testing and evaluation

There's no question that noise figure *measurement* is the most important test of converter performance and must be performed if proper operation is desired. As pointed out in reference 1, the measured noise figure may be optimistically low if a 5722 noise diode is used. The gas-discharge tube or hot-cold test are recommended above 400 MHz. The silicon diode noise generator is recommended only for optimizing the noise perform-



L1	12 turns no. 28 (0.3mm) on Amidon T25-12 toroid core
L2	7 turns no. 24 (0.5mm), air core, closewound on 0.1" (2.5mm) diameter
L3,L4	2 turns no. 22 (0.6mm), air core, 1/8" (3mm) diameter, 1/4" (6.5mm) long
Q1,Q2	Fairchild 2N5179 recommended but 2N2857, 2N918, FMT2060 or equivalent may be substituted

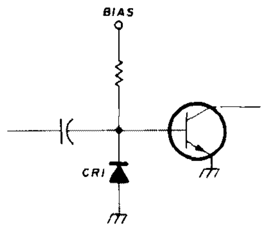
fig. 3. Two frequency doubler circuits which are recommended for use with vhf/uhf converters. The circuit in (A) is suggested for inputs in the range from 90 to 120 MHz. The doubler in (B) is recommended for inputs in the range from 180 to 220 MHz. The 1N914 diode in series with the power supply lead is the "idiot" diode discussed in the text.

ance of a converter – it can't be used for quantitative measurements.

The primary things to look for when evaluating converter noise figure are poor image rejection and nonlinearities in the i-f and detector. One good test is to compare your converter or preamplifier with another one (preferably one with a known noise figure) and note the difference. You can get a good idea of how well your converter measures up and absolute numbers are not necessary. For more information on the subject of noise figure, references 12 and 13 are recommended as starting points.

It's not easy to test for spurious responses but the simplest test is to substitute and/or bypass modules in

fig. 5. Simple limiter circuit is helpful in reducing burnout problems in preamplifier input stages. Diode CR1 is a low capacitance (1.0 pF maximum) silicon or hot-carrier diode such as the Hewlett-Packard 5082-2810.



tion of the undesired signal is usually necessary before corrective action can be taken. If you are fortunate enough to have access to a spectrum analyzer, spurious troubles can be quickly pinpointed.

If you use the modular approach, gain compression and intermod performance are easily checked without expensive test equipment. All you need is a coaxial type attenuator in the 10 to 15 dB range. If this is not available, 100 feet (30m) of RG-58/U coaxial cable is acceptable (at 432 MHz this represents approximately 12 dB attenuation).

The test procedure is to connect the attenuator between various stages of the receiver to see where the problem is being generated. If the attenuator is placed after the preamplifier and the problem persists, the preamplifier is probably the source of the difficulty. If the problem goes away, however, it's probably being generated further down the chain and a similar test of the mixer is required.

Once you have pinpointed the problem it can be dealt with accordingly. It may even be desirable to have a built-in system for switching in additional attenuation ahead of the mixer if the offending signal is not always present (such as a local television transmitter, etc).

Converter stability can best be tested by using a stable crystal calibrator with known stability. A variable voltage supply on various converter circuits can provide

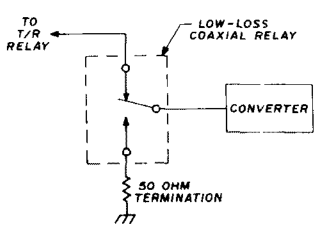
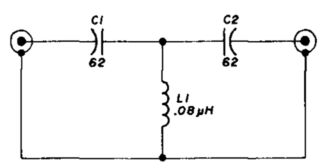


fig. 6. Simple relay protection circuit which enhances transmit-receive relay isolation.

fig. 7. Simple 50-MHz high-pass filter to decrease the effects of lightning and high-frequency interference. Keep all leads as short as possible to prevent circuit losses.



C1,C2	62 pF miniature capacitor
L1	0.08 µH. 4 turns no. 20 (0.8mm), air core, 1/4" (6.5mm) diameter, turns spaced one wire diameter

clues to the source of stability problems. A heat gun or hair dryer can be used to check for thermal instabilities.

Burnout is very difficult to evaluate since most burnout tests are, by nature, catastrophic. Suffice to say that if you have followed all of the previous recommendations, you have done about all you can do.

final remarks

The intent of this writeup is not to frighten you but to get you to try your hand at improving your own vhf or uhf converter. You will probably find that even if only one circuit is improved it will be worthwhile. Whenever you replace or modify a circuit you should *always* check the noise figure to make sure that the change resulted in an improvement – you may find that it was a move in the wrong direction!

I would like to once again reiterate that the use of quality components and the modular approach to vhf/uhf converter design will quickly pay for itself in terms of performance, versatility, and numbers of stations worked. It will also spur you on to trying new circuits and devices as they rapidly become available in today's highly volatile communications industry.

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ham radio

5/8-wavelength vertical antenna for two meters

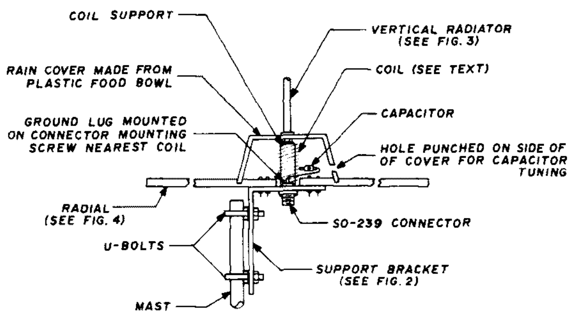
An easy-to-build antenna that provides a theoretical 3-dB gain over a quarter-wave groundplane

After operating two-meter fm mobile for some months I decided to operate from my home and built a vertical 5/8-wavelength antenna. Comparison with a 1/4-wavelength groundplane showed improved performance. I was able to work repeaters with the 5/8-wavelength vertical that I couldn't work with the groundplane. The antenna described here has been in operation since April, 1973. It's only 25 feet (7.6m) above ground. The lowest vswr obtained was 1.1 at 146 MHz, and the highest was 1.5 at 147 MHz. The antenna was made from junk-box parts and a discarded television antenna. Power levels of 100 to 150 watts can be handled safely. Greater power can be used by changing the series capacitor (**fig. 1**) to a variable capacitor with air dielectric, such as the APC type.

construction

Construction isn't critical and parts are easily obtained. The coil-support ceramic spacer and capacitor were obtained from a surplus military radio. The plastic weather cover was acquired from a local supermarket. You'll need an electric drill, drill bits as shown in the detail drawings, a 10-32 (M5) tap and holder, and a tapered reamer to make the 5/8-inch (16mm) diameter hole for the SO-239 connector. A bench vise is handy to bend the aluminum support bracket (**fig. 2**). The U-bolts were purchased from a local radio-TV parts dealer.

The assembled antenna is shown in **fig. 1**; details for making the support bracket, vertical radiator, and ground radials are given in **figs. 2** through **4**. Stainless steel hardware should be used if available. I used only three screw sizes to assemble the antenna: 4-40 (M3) for mounting the SO-239 connector, 6-32 (M 3/5) for mounting the radials, and 10-32 (M5) for the coil-support section. The 10-32 (M5) threaded stud at the top of the coil support was made from a 1½ inch (38mm) long screw with the head removed. A lug, lock



quantity	description
1	support bracket (reference fig. 2)
4	radials (reference fig. 4)
1	vertical radiator (reference fig. 3)
1	plastic cover (see text)
2	mast U-bolts (see text)
1	coaxial SO-239 connector
1	coil support — ceramic steatite pillar, Birnbach no. 445G or equivalent size 2½" (63.5mm) high by 3/4" (19mm) diameter threaded 10-32 holes both ends
1	capacitor — 7-45 pF ceramic trimmer, Erie type no. 503-000D7-45 or equivalent. Do not use mica compression type

Note A. Top of radiator can also be sealed by squeezing in vise, using a plastic cap of appropriate size, or filling the end with 2" (51mm) of caulking compound.

fig. 1. Assembled 5/8-wavelength vertical antenna for two meters.

washer, and nut are installed at the top of the support bracket and another lug on the screw nearest the coil support holding the SO-239 connector.

After assembling the parts described above, wind a coil of 9 turns using number 10 or 12 AWG (2.6 or 2.1mm) solid enamelled copper wire around the coil support. Space the turns to occupy the length of the support (2½ inches or 63.5mm). Solder the coil ends to the

By Ed Spadoni, W1RHN, 91 Tower Street, Dedham, Massachusetts 02026

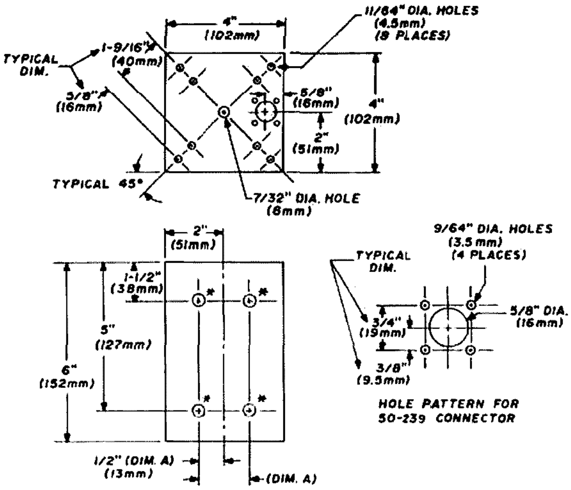
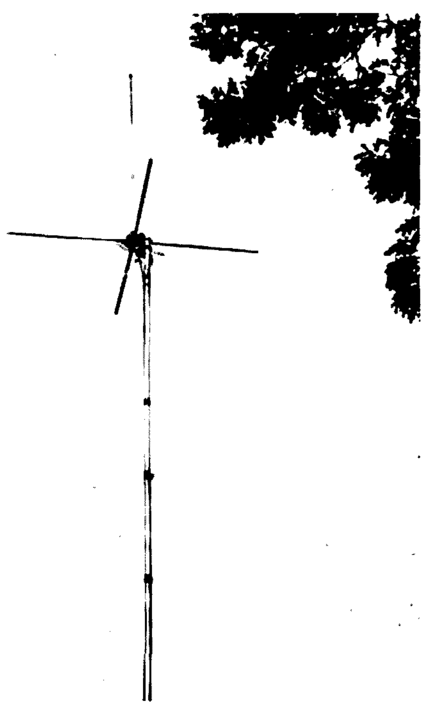


fig. 2. Support bracket details. Dimension A and holes marked with an asterisk are sized to fit available U-bolts. Material is 1/8" (3mm) aluminum.

lugs (see **fig. 1**). Next, solder the capacitor to the coil, two turns up from the ground end of the coil. Attach the other end of the capacitor to the center connector of the SO-239 connector. Install the capacitor so that the adjustment screw is accessible from the side of the cover.

To install the plastic rain cover, drill a 7/32-inch (6mm) diameter clearance hole on top for the stud to project through the cover. Slip the cover over the stud, and a flat metal washer and a lockwasher, then screw on the vertical radiator until it's tight. A capacitor-adjustment hole is made through the side of the plastic cover by aligning a scriber or centerpunch opposite the capacitor adjusting screw and penetrating the plastic cover. The hole should be about ¼ inch (6.5mm) diameter.

Installation of the two-meter vertical at W1RHN.



tuning

I used my ICOM IC-20 and a Bird vhf directional wattmeter to adjust the antenna for lowest vswr. With

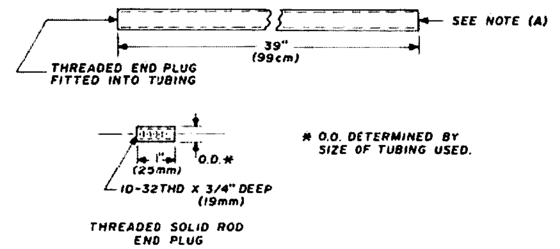


fig. 3. Construction details for the vertical radiator. Material is aluminum tubing, 7/16" (11mm) or 1/2" (13mm) OD, 1/16" (1.5mm) thick wall. Solid aluminum rod, if available, will provide sturdier construction and will avoid need for making end plugs to seal top. The end plug fits flush with the end of tubing and is held in place by "dimpling" with a center punch.

power applied to the antenna, I adjusted the capacitor with a nonmetallic alignment tool for the lowest reflected power on the wattmeter, using a test frequency of 146.52 MHz. If you don't have a vhf swr bridge or directional wattmeter, a field-strength meter can be used as an indicator. When using this method, adjust the capacitor for maximum output.

final comments

As with all vertical antennas, this 5/8-wavelength antenna is omnidirectional. However, it has approximately

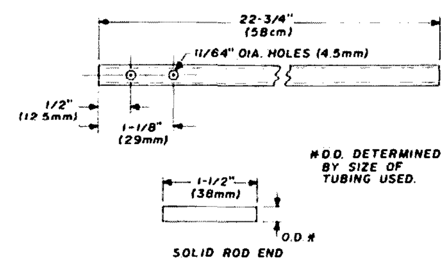


fig. 4. Radial element construction details. Install end plug flush with tubing, then drill 11/64" (4.5mm) holes ("dimple" plug in place before drilling).

3 dB gain over a ¼-wavelength groundplane and is an excellent antenna for all-around operation. With it I can consistently acquire repeaters 50 to 60 miles (80 to 96km) away. Under good conditions contacts 100 to 200 miles (160 to 320km) away have been made, both through repeaters and in the simplex mode.

ham radio

MICROPROCESSORS

8080 microcomputer output instructions

In the **microprocessor column** last month we discussed different types of simple input/output (I/O) devices and provided a listing of general principles of interfacing that apply to a wide variety of computers. This month, we would like to explain how software or computer instructions cause an I/O device to operate.

controlling power with a microcomputer

The I/O device that we shall choose for our discussion is the optically isolated solid-state ac relay. These relays can control any ac power device within the output current ratings of the relay. Shown in fig. 1 are typical solid-state relays which are available at prices ranging from $5 to $20 in small quantities. These relays permit a single TTL output signal of logic 0 or logic 1 to control up to 10 amperes of 220 Vac power (the Hamlin model 7522 relay shown at the top center of the photograph). Internally each relay contains a light-emitting diode, a light-sensitive transistor, a power triac, and a transparent dielectric optical path that isolates the digital and power circuitry and can withstand a voltage difference of at least 1000 volts.

A typical microcomputer I/O circuit that employs the solid-state relay is shown in fig. 2. Recall from the preceding column that the microcomputer sends synchronization pluses, called *device select pulses,* to the I/O device. In fig. 2 these are the pulses from the SN74154 decoder circuit. For an 8080 microcomputer each pulse from the decoder has a time duration of only 500 ns. It should be clear that a single 500 ns pulse cannot sustain the continuous operation of an ac power device. What is required is a simple interface between the microcomputer and the solid-state relay that will permit the ac power device to operate continuously, if it is so desired. A suitable interface is a single SN7474 positive-edge triggered flip-flop and a single buffer from a SN7407 hex buffer/driver chip. The buffer is needed since it is not good engineering practice to drive a solid-state ac relay directly from the output of a flip-flop.

With the aid of a suitable program, the microcomputer and SN74154 decoder can generate individual device select pulses that either clear or set the SN7474 flip-flop. To clear the flip-flop, and thus turn on the ac power device, only a *single* 500 ns pulse is needed. The flip-flop output, Q, will remain at logic 0 until a single 500 ns pulse is applied to the preset input, at which time the ac device will turn off.

It should be noted that any simple open collector gate or inverter can be used as the buffer between the output of the flip-flop and the input of the solid-state relay. Suitable choices would be the SN7401 or SN7403 2-input NAND gates, a SN7405 inverter, or a SN7409 2-input AND gate.

the output instruction

We shall discuss the subject of microcomputer instructions in considerable detail in subsequent columns. To summarize such discussions, there are 78 different instructions for the 8080 microprocessor chip, and a total of 256 variations of these instructions. Each instruction contains a single 8-bit *instruction code,* which indicates which type of operation or group of operations the microcomputer will execute. Some instructions contain two or three 8-bit bytes that are present in successive memory locations. A *byte* is defined as a group of eight contiguous bits occupying a single memory location.[1] Thus, 8080 microprocessor instructions are either 8, 16, or 24 bits long, with the first eight bits always being the instruction code.

The *out* instruction is a 16-bit instruction that consists of two successive 8-bit bytes located in successive memory locations. The first byte, in binary code, is always 11010011_2. The second byte can be any 8-bit binary number from 00000000_2 to 11111111_2 (the subscript 2 denotes a binary code); this is the device code of

By Peter R. Rony, David G. Larsen, WB4HYJ, and Jonathan Titus

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

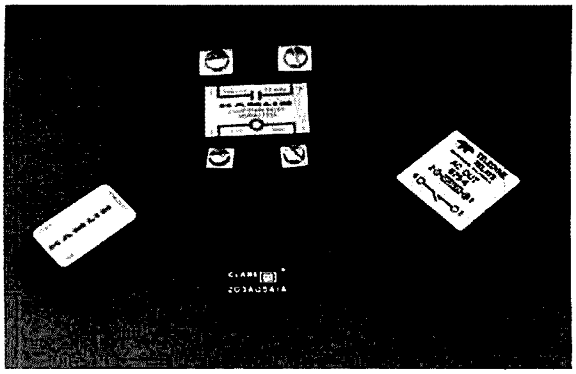
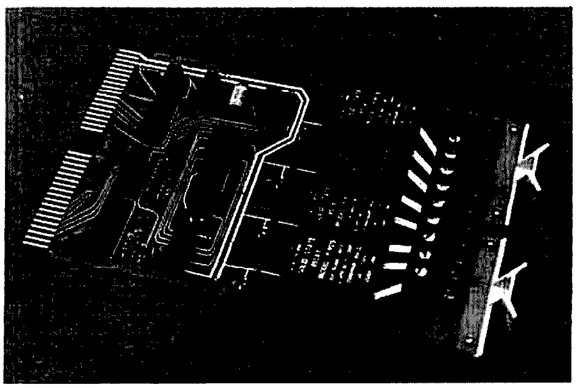


fig. 1. Typical optically isolated solid-state relays, including, from top left to top right, Hamlin 701-11-5 (1.5A, 120 Vac), Hamlin model 7522 (10 A, 220 Vac), Teledyne 657-4 (3 A, 120 Vac), and Clare 203A05AIA (750 mA, 120 Vac). The printed-circuit board shown at the bottom contains four Electronic Instruments and Specialty Corp. relays (3 A, 120 Vac) and all necessary external device addressing circuitry.



the specific output device that will receive eight bits of data from the accumulator. The instruction can be summarized as follows:

11010011_2 Generate a device-select pulse, with the aid of an 8-line-to-256-line decoder circuit, to allow an 8-bit data byte present in the accumulator to be sent to the desired output device. The contents of the accumulator remain unchanged.

simple program

The simplest program that incorporates the *out* instruction is probably the one given below:

memory address	instruction byte	description
0	11010011	Send device-select pulse to device given by following 8-bit device code
1	00000000	Device code for clear input to SN7474 flip-flop
2	01110110	Halt the microcomputer

An 8080 microcomputer operating at a clock rate of 2 MHz will execute this program in 8.5 μs. The ac power device will remain on once the program has been executed. To turn off the device, a slightly different program is required:

memory address	instruction byte	description
0	11010011	Send device-select pulse to device given by following 8-bit device code
1	00000001	Device code for preset input to SN7474 flip-flop
2	01110110	Halt the microcomputer

The ac power device will turn off after the second instruction byte in the program and remain off after the microcomputer halts. A more practical program requires additional instructions. Several such programs can be found in reference 1. Many of them have the following basic form:

memory address	instruction byte	description
.	.	
.	.	
.	.	
23	11010011	Turn solid state relay *on*, i.e., send device-select pulse to the device given by the following 8-bit device code
24	00000000	Device code for clear input to SN7474 flip-flop. When the flip-flop is cleared, the solid-state relay turns *on*

This section of the program may have various decision points that determine whether or not the solid-state relay is turned off. Typical decisions include

- Has sufficient time elapsed?
- Has the antenna reached its correct azimuth?
- Is the temperature of the final amplifier too high?
- Is the vswr too high?

.	.	
.	.	
.	.	
107	11010011	Turn solid state relay *off*, i.e., send device-select pulse to the device given by the following 8-bit device code
110	00000001	Device code for preset input to SN7474 flip-flop. When the flip-flop is set, the solid-state relay turns off.
.	.	
.	.	
.	.	

Keep in mind that a memory address contains sixteen bits. When we write "memory address 0" we really mean the memory address corresponding to the following 16-bit binary word: 00000000 00000000. Note that we have split the sixteen bits into two parts, the most significant eight bits and the least significant eight bits. These are called the HI (or H) and LO (or L) memory addresses, respectively.

With the aid of the above program, the solid-state relay shown in fig. 2 will turn on and off according to various decisions made by the program. A typical microcomputer-controlled system could easily have several such relays.

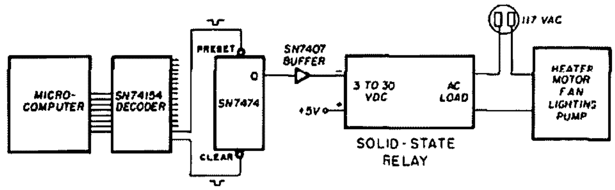


fig. 2. A typical I/O circuit for a power ac device such as a fan, heater, motor, or antenna control.

In a more orderly and systematic treatment of the 8080 microprocessor, you would probably introduce the 8080 instruction set prior to the discussion of any particular instruction, such as the *out* instruction described this month. Since we do not believe that you are willing to wait four months until we get to the *out* instruction, we have decided to treat it first. Next month we will provide a simple microcomputer program that generates device-select pulses to turn a device such as an antenna rotator or a fan on or off.

The authors will present a two-day seminar on microcomputers at the Virginia Polytechnic Institute and State University Extension Center in Reston, Virginia sponsored by Virginia Polytechnic Institute and State University Extension Division on March 12-13, 1976, and a five-day short course on digital electronics (with some discussion of microcomputer interfacing) sponsored by the American Chemical Society and Virginia Polytechnic Institute and State University, in Blacksburg, Virginia, on March 21-26, 1976. Two one-day seminars on microcomputers (sponsored by *ham radio magazine*) will be given at the Dayton Hamvention, Dayton, Ohio, on April 23-24, 1976.*

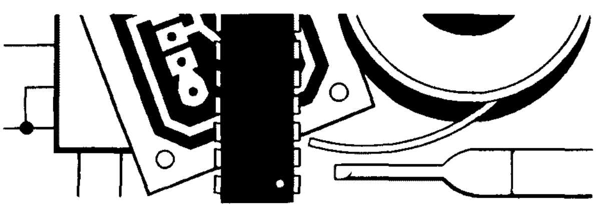
*The fee for the one-day seminar is $50 and includes $35 worth of books. Since the seminars are limited to 100 persons, early registration is recommended. For details write to *ham radio*, Greenville, New Hampshire 03048.

bibliography

1. David Larsen, Peter Rony and Jonathan Titus, "Bugbook III, Mocrocomputer Interfacing Experiments Using the Mark 80R Microcomputer, an 8080 System," E&L Instruments, Inc., Derby, Connecticut, 1975. $14.95 from Ham Radio Books, Greenville, New Hampshire 03048.

ham radio

the weekender



high-performance bench power supply at low-performance cost

In today's solid-state world, the variable low-voltage bench power supply is an absolute necessity for the serious experimenter. In fact, an axiom might well be that no matter how many bench supplies you have, you are always one short. And except for those amateurs who are fortunate enough to have acquired commercial power supplies, most of us are probably still using units which were built years ago and which go down to perhaps 4 or 5 volts because a reference supply would have been needed to enable the main supply to go to zero.

The relatively new RCA CA3130 operational amplifier makes possible a highly regulated power supply whose output will approach zero without a separate internal reference source. As shown in **fig. 1**, this op amp comprises both mos and bipolar transistors on a monolithic chip. It features a gate-protected input stage which has an input impedance of 1.5 *million megohms* and which is rated for typical input bias currents of 5 *picoamperes*[1]. It also operates from a single power supply. Best of all, the single-unit price is less than $2.00.

A power supply utilizing the CA3130 in a regulator circuit described in reference 1 has been constructed. It provides a regulated output of 0 to 40 volts at better than one ampere, and incorporates fold-back short-circuit protection. Load regulation is 0.1 percent, and line regulation is 0.02 percent for a 10-percent line-voltage change. Total noise and ripple output is less than 200 microvolts rms.

The CA3130, four discrete transistors, an inexpensive transistor array IC, and a bridge rectifier constitute the entire semiconductor complement, at a total cost of under $11.00. Furthermore, all of the devices are available from your friendly RCA Solid State distributor, if you specify the equivalent RCA type 44002 diodes in lieu of the 1N4002s indicated in fig. 2*.

*A complete parts kit for this power supply is being made available in conjunction with this article. For ordering information and prices, write to Dentron Radio Co., Inc., 2100 Enterprise Parkway, Twinsburg, Ohio 44087, or telephone (216) 425-3173.

By Robert S. Stein, W6NBI, 1849 Middleton Avenue, Los Altos, California 94022

circuit description

The complete circuit of the power supply is shown in **fig. 2**. In this unit, T1 is a surplus transformer having a 36-volt, 2-ampere secondary. However, you can use any transformer you wish, up to 40 volts, by selecting the values of two resistors (more about this later).

The operating and reference voltages for U2, the CA3130, are established by U1, a CA3086 transistor array, which is shown in block form to simplify the schematic. A stable, low-impedance, temperature-compensated source of reference voltage is provided at pin 14 of U1. Voltage adjust control R7 determines the output voltage by setting the amplitude of the reference voltage which is applied to the inverting input (pin 2) of U2. In turn, the output of U2 supplies the base of driver transistor Q3, which controls the base current of the Darlington pair (Q1 and Q2), and hence the series pass resistance.

Let's assume that the output from the power supply starts to increase, either because of an increase in ac line voltage or a decrease in dc load current. A proportional increase appears at the junction of resistors R13 and

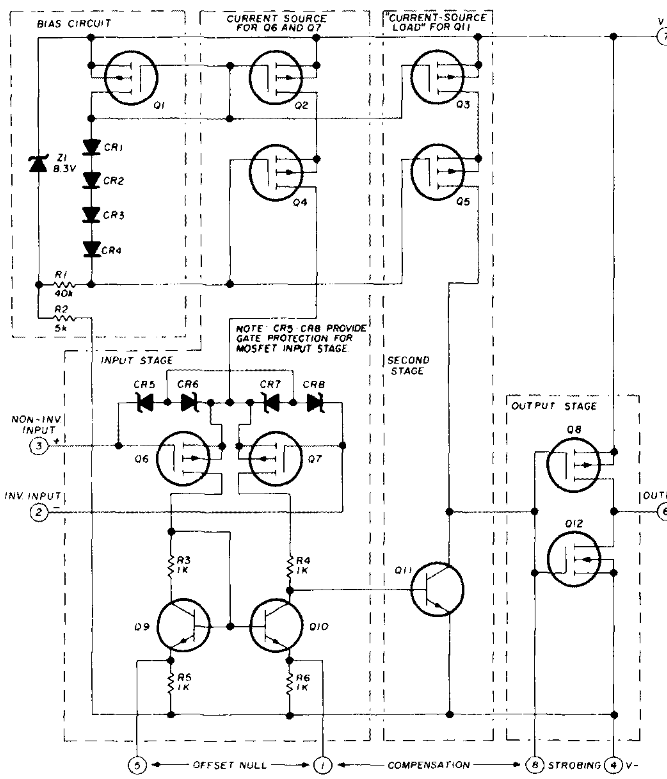


fig. 1. Schematic diagram of the RCA CA3130 op amp.

R15, and is applied to the non-inverting input (pin 3) of U2 through R14. The resultant increase in output current from U2 causes an increase in the collector current of Q3, reducing the drive to the base of Q2 and increasing the collector-to-emitter resistance of Q1. Thus the output voltage is reduced by the closed loop until it is reestablished at the value set by R7. The entire correction takes place almost instantaneously, abetted by the high gain of the op amp (approximately 110 dB) and of the loop. Conversely, if the output voltage starts to decrease, an equivalent but opposite reaction maintains the output constant.

Short-circuit protection is provided by Q4. The base-to-emitter voltage of Q4 is the voltage which appears across the base-emitter junction of Q1 plus the voltage drop, caused by the load current, across parallel resistors R8 and R9. When this voltage exceeds a predetermined value established by potentiometer R11, Q4 conducts and diverts current from the base of Q1, thereby limiting the short-circuit current to a safe value (approximately 500 milliamperes) and saving the series-pass transistor.

To quickly reduce the output voltage to zero when primary power is switched off, a double-pole double-throw switch is used. One pole of S1 is the conventional primary ac switch; the other pole discharges the filter capacitors through R16 when the switch is in its *off* position. A 50-volt meter is incorporated to monitor the output voltage.

Returning to the function of U1, let's examine the configuration of its transistors. Fig. 3A shows the actual circuit arrangement, while fig. 3B shows the functional circuit. Transistors Q_A and Q_B are each connected to form a zener diode, and are connected in series to provide a regulated 14-volt supply for U2.

Transistor Q_E is configured as a constant-current

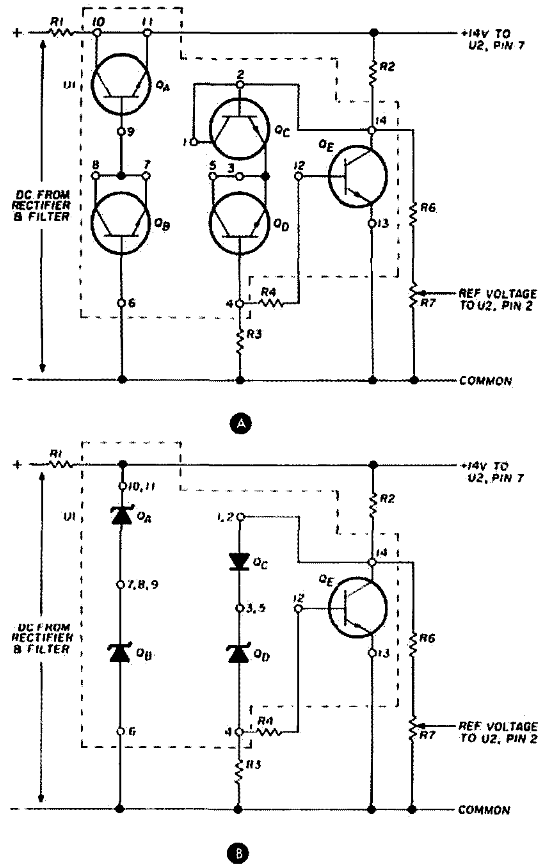


fig. 3. Circuit A is the actual dc circuit arrangement of transistor array U1, as used in fig. 2. Circuit B is the equivalent functional circuit. The numbered terminals correspond to the pin-outs on the CA3086 package.

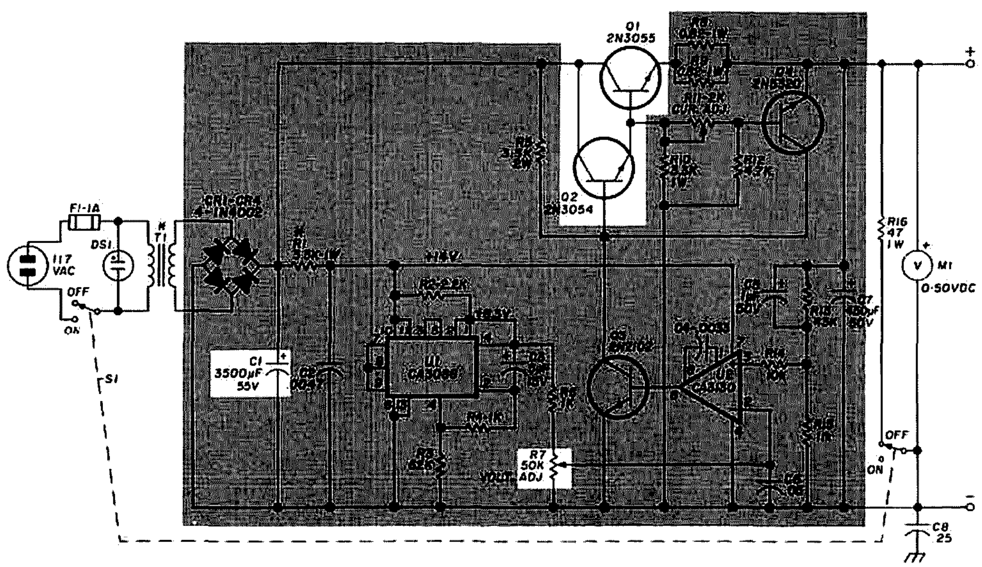


fig. 2. Schematic diagram of the regulated power supply. Parts shown within the shaded area are mounted on the printed-circuit board. Those marked with an asterisk are discussed in the text.

generator whose base bias is dropped from the collector potential by Q_C and Q_D. The latter is connected as another zener diode, while Q_C provides temperature compensation by virtue of its arrangement as a forward-biased silicon diode. This entire circuit, operating from the regulated 14-volt line, establishes a stable reference-voltage source of approximately 8.3 volts at the collector of Q_E.

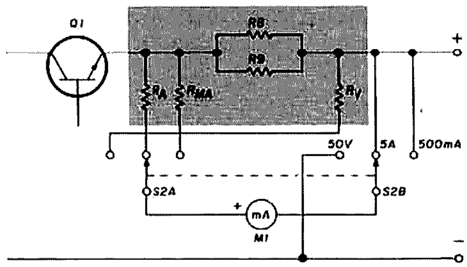


fig. 4. Voltage and current metering circuit using a single milliammeter. The values of R_A, R_{MA}, and R_V depend on the meter, and their determination is covered in the text. The switch *must* be a *non-shorting* type. Q1, R8, and R9 are the same as in fig. 2. Parts within the shaded area are mounted on the printed-circuit board.

A portion of this voltage is applied, via R6 and R7, to the inverting input of U2 and serves as the reference for the regulator circuit.

transformer

Since it is unlikely that anyone else would have a transformer identical to the one I used in my supply, a brief discussion of the transformer is warranted. Obviously, the first requirement is that the secondary be rated for at least the maximum amount of current you expect to draw; one ampere is recommended.

The secondary voltage can by anything between 15 and 40 volts. The choice depends on the maximum dc output that you want, which will be about 5 volts less than the unregulated dc developed across filter capacitor C1. A good transformer will supply unregulated dc equal to 1.4 times the secondary rms voltage at no load, and about 1.3 times the secondary rms voltage at full load.

Using these figures as the starting point, calculate a value for resistor R1 which will drop the *estimated no-load dc voltage* to 14 volts at a current of 10 milliamperes. R15 will probably also require a change from the value shown in **fig. 2**, but this can best be done after the power supply is built and working.

voltage and current metering

Although only the voltmeter shown in **fig. 2** is really necessary, it is often convenient to be able to measure the load current without having to hook up an external meter. Fig. 4 shows a simple circuit using a milliammeter and a three-position switch to allow measurement of output voltage and measurement of load current in two ranges. In order to keep the total current from passing through the switch contacts, the voltage drop across the overload-sensing resistors (R8 and R9) is measured, with the meter indicating the equivalent current through the resistors. The values or R_A, R_{MA}, and R_V depend on the ranges desired, the full-scale meter current, and the meter resistance; the method of calculating these values is covered in the appendix.

construction

The housing and construction of the power supply is a matter of personal preference. Naturally, for bench use, switch S1, pilot light DS1, meter M1, voltage control R7, the output terminals, and meter switch S2 (if used) should be on the front panel. Current adjust potentiometer R11 can be a screwdriver-adjust or pc-board type, since it is set once and then forgotten.

Transistors Q1 and Q2 must be mounted on a husky heatsink – the larger the better. Remember that Q1 passes the full load current, and dissipates power equal to this current times the voltage drop between collector and emitter. This can exceed 40 watts at low output voltages, which is a lot of heat to dissipate. Be sure to use insulators between the heatsink and the transistors,

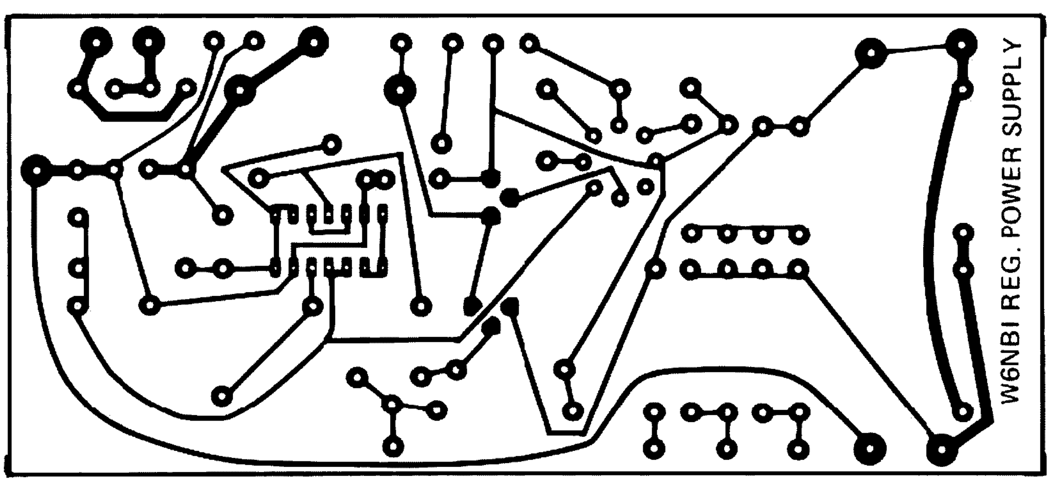


fig. 5. Foil pattern of the printed circuit board.

and to apply a thin film of silicone heat-transfer compound to each side of the insulator.

Most of the other parts are mounted on the printed-circuit board shown in figs. 5 and 6, or can be hard-wired on a piece of perf board. The pc board has been designed to accept standard available parts. R11 is a thumbwheel-type control, such as a Mallory MTC23L1 or CTS X-201-R202B. The 2.5k Radio Shack 271-228 may also be used.

maximum output voltage cannot be reached, R15 must be decreased. In either case, small changes in R15 will permit full rotation of the voltage adjustment control over the desired voltage range.

Reduce the output voltage to zero and connect a load resistance to the power supply (in series with an external ammeter if you have incorporated only the voltmeter) which will draw approximately one ampere at any output voltage over 5 volts. Monitor the load current and

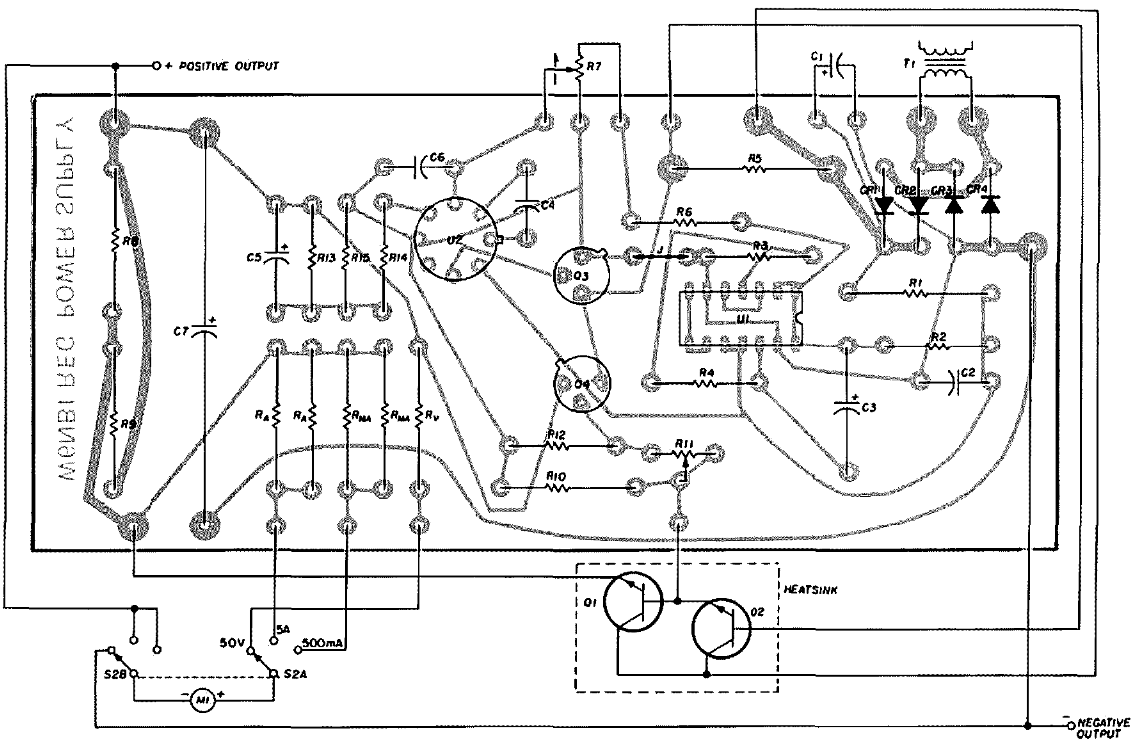


fig. 6. Component layout of the printed-circuit board. R_A, R_{MA}, and R_V are part of the optional metering circuit shown in fig. 4, and may be omitted if that circuit is not used. Note that provisions have been included for incorporating two parallel resistors for both R_A and R_{MA}, as discussed in the text. J signifies a wire jumper.

If you use the perf board, remember one thing: be certain to connect the negative terminal of filter capacitor C1 directly to the negative side of the rectifier bridge. None of the other negative returns is critical, but must not be between the capacitor and the rectifier. Failure to wire this or any other high-current power supply in the prescribed manner will result in excess ripple in the output.

adjustment and test

After all wiring has been checked, set voltage adjustment control R7 to minimum and current adjustment potentiometer R11 to mid-range. Apply ac power and monitor the output voltage while rotating the voltage adjustment control to maximum. The output voltage should increase smoothly. If it reaches a maximum before the control is fully rotated, the value of R15 must be increased. On the other hand, if the desired

gradually increase the output to one ampere. If you cannot obtain enough current, readjust pot R11 slightly. Check the output voltage as the load resistance is disconnected and reconnected; there should be no discernable voltage change.

Once more reduce the output voltage to zero and set R11 for minimum resistance between the base of Q1 and the base of Q4. Disconnect the load resistance and short-circuit the output terminals if you have built the ammeter circuit into the power supply. Otherwise connect an external ammeter (1 amp or more) directly across the output terminals. Slowly increase the setting of the voltage adjustment control, and adjust R11 for a short-circuit current between 450 and 500 milliamperes.

Finally, check the value of R1 by measuring the voltage across it under no-load conditions, and calculate the current flowing through it. The calculated current should be approximately 10 milliamperes. If it is over 11

or less than 9 milliamperes, change R1 to obtain a value of current closer to 10 milliamperes.

That completes the project. If you want to check the regulation, you will need a digital or differential voltmeter, since a 0.1-percent change is measured in millivolts. Noise and ripple output can be checked with either a high-sensitivity scope or with a good ac electronic voltmeter, such as a Hewlett-Packard 400D, capable of reading 1 millivolt full scale. If you have test equipment of this type, have at it. Otherwise, forget about it and make a new addition to your bench – you can always use one more power supply.

appendix

Calculation of resistance values for R_A, R_{MA}, and R_V in **fig. 4** must be based on the following known factors: the exact parallel resistance of R8 and R9, the internal resistance (R_m) of the meter, and the full-scale meter current (I_m). The meter characteristics generally are known or can be measured. However, the parallel resistance of R8 and R9 must be accurately known or measured on a bridge (not calculated from the nominal resistance values), or an external ammeter must be used for calibration.

The value of R_V can be calculated from the equation

$$R_V = \frac{E_{fs} - (I_m R_m)}{I_m}$$

where E_{fs} is the desired full-scale voltage.

The value of R_A (or R_{MA}) is determined by the equation

$$R_A = \frac{(I_L R_{8\text{-}9}) - (I_m R_m)}{I_m}$$

where I_L is the load current corresponding to full-scale meter current and $R_{8\text{-}9}$ is the parallel resistance of R8 and R9.

As an example, assume that the meter ranges in **fig. 4** are to be obtained with a 1-mA meter having an internal resistance of 100 ohms.

Using the preceding equations,

$$R_V = \frac{50 - (.001 \times 100)}{.001} = 49{,}900 \text{ ohms}$$

$$R_A = \frac{(5 \times .41) - (.001 \times 100)}{.001} = 1950 \text{ ohms}$$

$$R_{MA} = \frac{(.5 \times .41) - (.001 \times 100)}{.001} = 105 \text{ ohms.}$$

The meter will be sufficiently accurate if one-percent resistors are used to obtain these values. However, if the actual resistance of R8 and R9 in parallel was not known, and the nominal value was used to calculate the values of R_A and R_{MA}, the current ranges may be off by 5 to 10 percent, depending on the actual resistances of R8 and R9. This can be improved if a more accurate ammeter is available. Instead of using the calculated values of R_A and R_{MA}, use one-percent resistors which are approximately ten percent higher than the calculated values. Then connect the external ammeter in series with a load and compare the currents read on the external and internal meters. You can expect the internal meter to read low. Increase its reading by connecting a composition resistor in parallel with R_A (or R_{MA}). Start with one that has about ten times the resistance of the one-percent resistor, and keep trying different values until the two meters agree.

You can select any meter ranges that you feel are convenient, and then change the meter scale accordingly. I have found that the markings on most scales can be erased with a pink (not white grit) pencil-type typewriter eraser. New numbers can then be applied, using rub-on transfers.

reference

1. Data Sheet for CA3130 Series COS/MOS Operational Amplifiers, File Number 817, RCA Solid State Division, Somerville, New Jersey 08876, August, 1974.

ham radio

the ham notebook

another look at the fm channel scanner for the Heath HW202

The scanner adaptation article by Ken Stone, W7BZ, in the February, 1975, issue of *ham radio* was a good one. (The original article, by K2ZLG, appeared in the February, 1973, issue.) I'd like to add some words on my experience with this circuit.

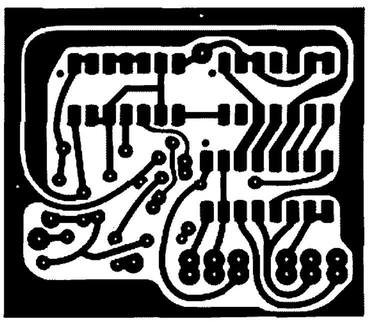


fig. 1. Circuit-board layout for the f-m scanner.

I wanted to have the scanner and the tone-burst encoder mounted *inside* the HW202 without the inconvenience and add-on appearance of outboard hardware. So I made the circuit board shown in fig. 1 for the scanner, which fits inside the HW202 just under the tone-burst encoder and just above the speaker.

Parts are hard to get in this area. The 7445 was replaced with a 7441, and all outputs were connected so the unit would scan in both directions. The 2N4140s were replaced with 2N2222s. The LEDs were cemented into small holes punched on either side of the tone-burst encoder knobs. Two of the same colored wires, about 8 inches (20.3cm) long, were soldered as close as possible to the LEDs.

A notch was made on the upper side of the hole where the tone-burst encoder shaft passes through the metal plates behind the knobs. This notch allowed enough room for the wires to pass over the switches and through the holes without binding. With the tone-burst encoder remounted, an area about 2-1/8 by 1-3/4 inches (54 by 45mm) was available for mounting a PC board for the scanner.

All scanner components were first mounted on a breadboard and all wiring was connected without cutting off any excess. This broadboard version was a good idea, because initial checks indicated some problems – the LEDs didn't agree with the channels below them. Make sure you check out the circuit before you button it up.

The PC board shown in fig. 1 looks somewhat unorthodox, but when nobody is around to tell you how, you do the best you can. Note that the letter J appears several times in the PC-board illustration. This means that a jumper must be connected between the points shown. Make sure you install the jumper under the 7441 before you mount the device, otherwise you'll have to install an insulated jumper on the foil side. Also note that pin 10 must be clipped from the 7490 or it can't be mounted. Finally, leave wire lengths longer than necessary to facilitate knob removal of the tone-burst encoder.

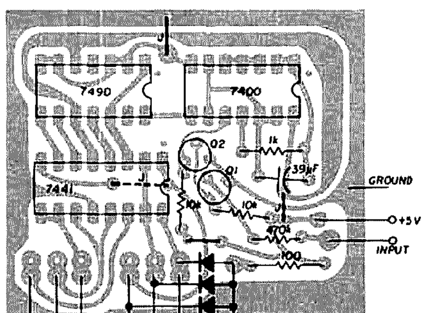


fig. 2. Component layout for the fm scanner board. Letter J designates a jumper.

Bill Biser, K7PYS

variable, low-cost power supply for transistor work

If you like to work with transistors and ICs, troubleshoot transistor radios, receivers, walkie-talkies, or are tired of buying batteries for low-power experimental work, this variable power supply may be the answer.

As this is a simple, low cost, one-evening project, only the essential features are described. You may want to add refinements of your own which will make the power supply more responsive to your own specific needs. Although I made no measurements of the output regulation, it appears adequate for the intended purposes.

The heart of this power supply are ac adapter units which are also known as battery eliminators, solid-state dc power supplies, power converters, etc. These adapters can be found at flea markets, hamfests and junk shops at bargain prices. I paid fifty cents each for mine. One unit alone could have been used in the circuit of fig. 3 although two are preferable for 12 or 18 volts output.

The two adapters I used had the following information imprinted on the cover:

120 Vac	9 Vdc, 40 mA
60 Hz	or
5 W	6 Vdc, 230 mA

Since I wondered how you got 6 V dc at 230 mA from a sealed unit whose

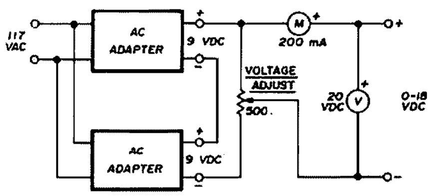


fig. 3. Simple, low-current power supply is based on the use of ac adapters or battery eliminators which are often found at flea markets.

actual output was 9 Vdc, I wrote to the manufacturer. They replied that "This power converter is intended to be used with our products at these voltages and current ratings." From this I concluded that the unit was capable of giving 6 volts at the prescribed current but that I would have to supply my own voltage-dropping resistor. All of this is not too pertinent except to point out that you should read the manufacturer's label and be sure to measure the voltage output. The important specification is the wattage rating of 5 watts – this is adequate for a low-current power supply.

These ac adapters usually consist of a small step-down transformer, full-wave rectifier (often selenium) and capacitor input filtering. The circuit is often on a printed-circuit board and is contained in a tightly sealed plastic box.

There is nothing critical in power supply layout or wiring. Be sure to insulate the binding posts from the chassis, at least the plus lead. In addition to the two ac adapters you need two panel meters, a toggle switch, a 500-ohm potentiometer, duplex wall outlet for 117 Vac, two binding posts, and an ac plug and line cord. The meters, switch and binding posts are mounted on the front panel. The duplex wall outlet is installed on the chassis from underneath. Since most of the ac adapters have plugs integrally molded into the housing, they can be plugged directly into the duplex outlet on the chassis. The output of the power supply I built is about three-quarters of a watt with voltage control from zero to 18 Vdc.

A word about the meters. Although I used 50 Vdc and 300 mA meters I had in my junk box, a 0-20 Vdc voltmeter and a 0-100 or 0-200 milliampere meter would be preferable.

Howard Stark, WA4MTH

microwave frequency doubler

As spectrum space becomes more and more valuable, amateurs are forced to explore the communication possibilities of the microwave bands. Klystrons and other exotic devices have been available since the late forties for use as high as 12 GHz with modifications, but there are few components available for the next higher amateur band (24 GHz), either surplus or commercial. This article describes a simple frequency doubler to get from 12 GHz to 24 GHz in one noncritical step. The step is made with an ordinary point-contact diode, a 1N23.

As most amateurs are aware, if a

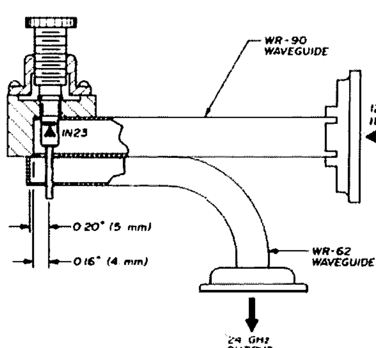
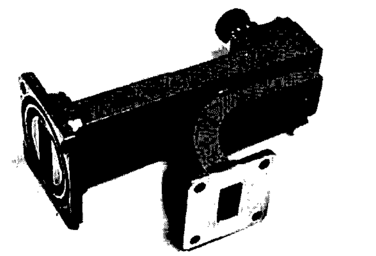


fig. 4. Microwave 12 to 24 GHz frequency doubler uses 1N23 point-contact diode. Only critical dimensions are spacing of the diode from the shorted ends of the waveguides.



Microwave frequency doubler provides 5 Mw output at 24 GHz with 100 mW input at 12 GHz. Fundamental is suppressed 23 dB.

diode is driven by an alternating current source, harmonics will be generated but the strength falls of with increasing harmonic number. Here only the second harmonic is used which is 13 dB down from the fundamental. The third harmonic is an additional 20 dB down so it can be ignored (or filtered out if you are a purist).

The diode cartridge is mounted in an untuned waveguide section as illustrated in fig. 4. The input waveguide is a short section of WR-90 guide, 1 by 0.5 inch (2.54 by 1.22cm) nominal. The output guide is type WR-62, 0.7 by 0.35 inch (1.78 by 0.89cm) nominal. By the way, the WR classification gives the largest inside dimension of the guide. For example, the inside width of WR-90 is 0.9 inches; the inside width of WR-62 is 0.622 inches.

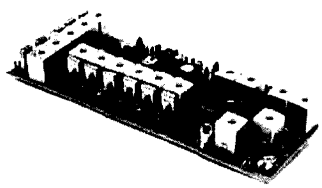
The only dimensions which should be adhered to closely in the construction of the doubler is the spacing of the diode from the shorted ends of the waveguides. The tolerance here is ±0.04 inch (±1mm).

In use, an 12-GHz klystron, such as a X-13, is connected to the input flange and 24-GHz energy will exit the output port. As previously mentioned, the system loss for a drive level of 100 mW is 13 dB. The fundamental is down 23 dB. The diode presently being used has withstood 350 mW of drive for over 200 continuous hours with no degradation in performance.

John Franke, WA4WDL

NEW products

vhf fm transmitter strip and power amplifier



The new T40 fm transmitter strip from Hamtronics, Inc., features crisp, clear, symmetrical modulation, is compatible with carbon or transistorized dynamic microphones and has separate microphone gain and deviation controls for proper modulation setup. It also has sufficient shielded tuned circuits to minimize harmonics and spurs.

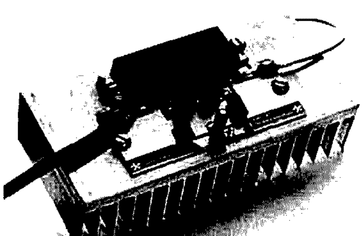
The exciter module is designed primarily for two meters; a tripler/driver is available for 450 MHz; other frequencies such as 50, 220 MHz, and the commercial frequencies are available on request. Output power is adjustable up to 200 mW with a power control pot. This level is sufficient to drive the companion power module, and it is also useful as a QRP signal around town or for repeater input. The T40 is also well suited for use as a multi-channel fm signal generator or as a control or supervisory link. The exciter module may also be used on CW for the low ends of the 144 and 432 bands.

The T40 has eleven channels, with ground-on diode switching suitable for busing with receiver control lines if desired. Sockets are provided for standard, series-resonant 12-MHz HC-25/U crystals, which are readily available. Separate, multi-turn vernier coils allow individual adjustment of each channel frequency.

The assembled G-10 PC board measures 3x7-1/2x1 inch (7.6x19x2.5cm). It is designed to slide into vertical grooves in the companion cabinet. It may also be standoff mounted in your own package or mounted on the rear of a rack panel. The unit is powered by +13.6 Vdc at 70 mA.

The new T80 rf power modules employ the new "rf modules" recently introduced by TRW and others. These bricks are like magic compared to the tricky and sometimes unstable discrete power amplifiers commonly in use today. The bricks are self-contained, take 200 mW drive and amplify up to a level of 20-25 watts on 2 meters or 13-15 watts on 432-450 MHz. There are no external tuned circuits to fuss with. There isn't any tuning at all! Believe it or not, you simply connect the exciter and antenna to the brick, add 13.6 Vdc, and you're on the air with instant power.

Hamtronics has packaged the bricks with the necessary decoupling components, connection facilities and heatsink and sell the unit all assembled and tested so you have no worry about what you are getting into. Features include vswr protection, no tuning, stability under all normal conditions including vswr, clean output signal, low power drain, and easy installation. The T80 is especially well suited for repeater service since it is unaffected by changes in antenna impedance due to weather.

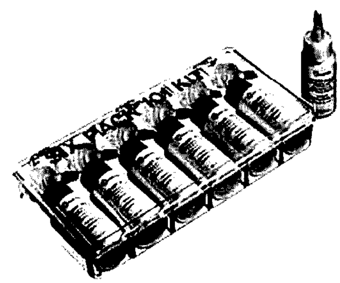


Input power is 13.6 Vdc at 2-4 amp, depending on drive level and frequency band. Efficiency is 30-50%. Connections for power and rf signals are provided through a PC board which butts up to the leads of the rf power module. Both are mounted to a heavy heatsink. The heatsink may be attached to the rear panel of the companion cabinet with the power amplifier mounted inside the rear panel, if desired, or it may be mounted to suit your installation. All you need do is attach your coax cables and power lead and you're on the air.

For uhf operation, the T20 uhf tripler/driver module is available to interface between the T40 exciter and the uhf power module. Housed on a PC board, it requires about 200 mW of drive at 2 meters to provide 200 mW output at 432-450 MHz.

Price of the T40 exciter module kit is $39.95. The T80 rf power module is $79.95, wired and tested. A companion cabinet is available for $24.95. The T20 tripler/driver module kit for 450 MHz is $19.95. For complete information, including an illustrated catalog, send an SASE to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

instant-weld adhesive



Need a third hand when building equipment? Oneida's new Six-Pak of industrial-strength alpha cyanoacrylate adhesive is available in one-ounce dispensers. Up to 11,400 one-drop applications are possible from this new package of dispensers.

Standard Six Paks come with an assortment of five Instant-Weld formulas; however, you may specify the assortment of your choice. The five formulas are designated by label colors: red, blue, yellow, green, and white. Red label sets the fastest. It bonds many materials in about three seconds, including plastics, rubber, ceramic, or glass.

Thousands of items and materials can be quickly and permanently bonded with Instant-Weld. If you have an unusual bonding problem, write Oneida Electronic Manufacturing, Inc., attention Mr. Dalton Smith, 138 West 7th

More Details? CHECK—OFF Page 102

Street, Erie, Pennsylvania 16501, or call (814) 336-2125. Oneida will make recommendations, which may require a special mix. Instant-Weld is made in the U.S.A. and is guaranteed. The standard Six-Pak sells for $49.95 postpaid. A brochure is available from Oneida Electronic Manufacturing, Inc., Meadville, Pennsylvania 16335 or use *check-off* on page 102.

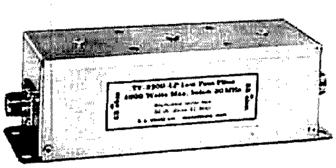
QSL bureau for novices

Here's good news for novice amateurs who would like to save money on sending and receiving QSL cards. Jim Isham, W8TX X, operates a novice QSL bureau that works like this: For only $2.00 per year, a novice may send him as many QSL cards as he wishes, and Jim will remail them. All novice amateurs are asked to keep a self-addressed stamped envelope on file with W8TXX to expedite the remailing system.

When a novice subscribes to Jim's service, Jim will send the novice a card with a code number, which must be included in all future mailings. The code is different for each subscriber. At the end of the year's subscription, the novice subscriber will be sent another card, which tells him the year is up and asks if he wishes to subscribe to the service for another year.

Considering the present postal rates for sending cards individually this service looks good indeed. If postal rates increase, savings will be even more. Write to Jim Isham, W8TXX for further details. His address is Box 1111, Benton, Harbor, Michigan 49022.

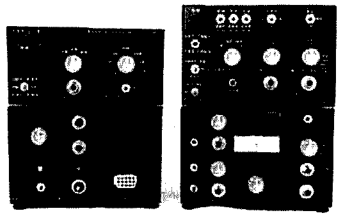
lowpass TVI filter



Television interference, even from modern transmitting equipment, is still a problem in many areas. A well-designed lowpass filter properly installed in an amateur transmitter can prevent or substantially reduce television

interference. The R. L. Drake Company has added a new lowpass filter, the TV-3300-LP, to their line of television interference filters. The TV-3300-LP provides more than 80 dB attenuation above 41 MHz and will handle 1 kilowatt maximum input below 30 MHz. More information? Write R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342, or use *check-off* on page 102.

avionic test-equipment kits



Radio Systems Technology has announced a new line of avionic test equipment. This equipment, designed to give the ham-pilot the capability of bench service of VHF Nav Com transceivers and Marker Beacon receivers, is available for the first time anywhere in kit form. Four separate kits allow the test bench to power 12-volt radios and accurately synthesize navigation and communication signals as well as measure transmitter and audio outputs. All kits may be powered from 110 volts ac, 12 volts dc or internal 9 volt batteries, and are supplied with rod antennas for use as ramp test sets.

The RST-721 Communications Test Set is designed to check all inputs and outputs of vhf communication transceivers for correct operation. This unit is a composite rf and audio wattmeter, voltmeter, ammeter, percentage modulation meter, vswr meter, microphone test set and crystal-controlled 122.8 MHz signal generator. Rf and audio 10-watt dummy loads are built-in. The RST-721 Test Set sells for $98.50.

The RST-711 Navigation (VOR-LOC) Test Set is designed to test vhf navigation receivers for correct operation. This unit is a crystal-controlled 108.0 and 108.1 MHz signal generator modulated with four major OMNI signals or LOC signals together with pulsed

More Details? CHECK–OFF Page 102

or continuous IDENT tone. A front panel rf level attenuator allows receiver sensitivity to be checked. The unit sells for $97.50.

The RST-701 Marker Beacon Test Set is a crystal-controlled 75 MHz signal generator modulated with any of three marker tones, Fan, Outer or Middle Marker. A front panel attenuator selects the desired rf level for checking sensitivity. The RST-701 sells for $69.50.

The RST-601 is a 12 volt bench regulated power supply which has the capability of simulating battery high limit voltage, nominal, and low limit voltage. Also included on the RST-601 chassis are microphone and headphone jacks, built-in speaker, and three parallel connectors for transceiver interconnection. The unit sells for $85.00.

The RST-601, 701, 711 and 721 may be purchased as a complete Bench Test Set for $330.00. For further information, write to Radio Systems Technology, P.O. Box 23233, San Diego, California 92123, or use *check-off* on page 102.

precision frequency comparators

The Dynatron Company has announced two frequency comparators intended for use in calibrating crystal oscillators against television network atomic standards. The comparator generates a vertical rainbow bar on the screen of a color television receiver. The rate of color change of the bar indicates the phase difference between the crystal oscillator output and the network color subcarrier. Frequency calibrations to within a few parts in 10^{10} take just a few minutes.

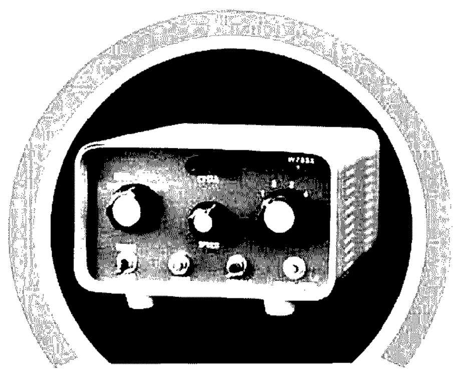
This calibration scheme, based on techniques and circuits developed by the National Bureau of Standards, is supported by the monthly publication of the network offset frequencies in the NBS Services Bulletin.

The DyCo Model 175 accepts input of 2.5, 5 or 10 MHz while the Model 175-1 accepts 1, 5 or 10 MHz. Price of the Model 175 is $99.95 while the Model 175-1 sells for $109.95. For more information contact the Dynatron Company, Post Office Box 48822, Los Angeles, California 90048, or use *check-off* on page 102.

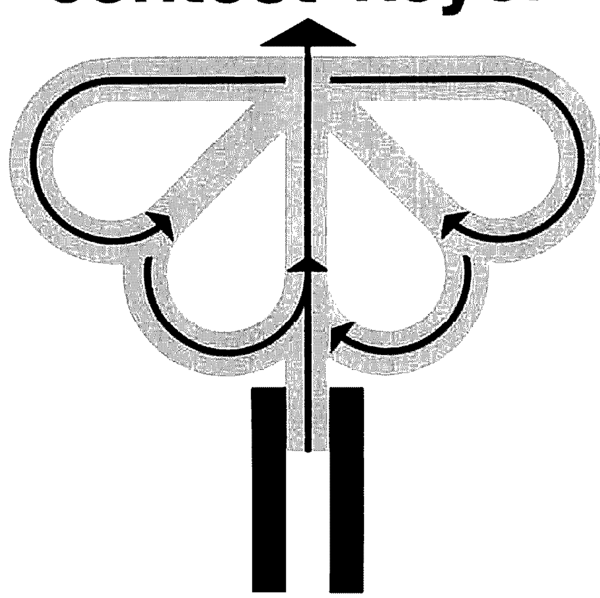
one dollar

ham radio magazine

APRIL 1976



programmable contest keyer



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offices
Greenville, New Hampshire 03048
Telephone: 603-878-1441

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Ham Radio Europe
Box 444
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Ham Radio France
20 bis, Avenue des Clarions
89000 Auxerre, France

United Kingdom
Ham Radio UK
Post Office Box 64, Harrow
Middlesex HA3 6HS, England

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Greenside, Johannesburg
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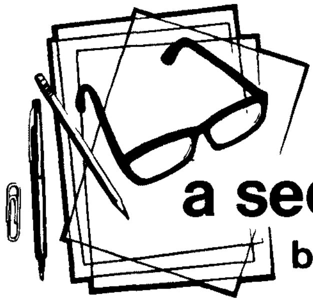


a second look

by Jim Fisk

Now that 16-kilobit random-access memories are starting to appear on the market, it shouldn't be too long before we see some of these devices in amateur products. Its predecessor, the popular 4k RAM, evolved rather slowly because manufacturers were forced to switch from p-channel to n-channel designs to reduce cell size. Since 16k RAMs use the same basic technology, prices can be expected to race down the curve at a much faster pace -- some manufacturers are predicting the price will drop to less than $10 by early next year. Others see a much slower pace, with the magic $10 price at least three years away. All agree, however, that once it's in production, the 16k RAM will dominate the solid-state memory market well into the early 1980's.

One of the reasons the 16k memory will be with us for awhile (as opposed to the relatively short-lived 1k, 2k and 4k devices) is because the next level of RAM integration, 65,536 bits, is probably beyond the reach of the n-channel MOS process. In the 16k RAM each bit is squeezed into a site about one-thousandth of an inch square (0.025mm^2) – about half the area required in 4k designs – by placing the cell's switching transistor and storage capacitor on top of one another instead of side by side as in the 4k layout. However, most researchers are convinced that the switched-capacitor or single-transistor cell used in present RAMs won't be good enough for 65k devices – it will probably peter out well short of 0.3 to 0.5 mil^2 (0.0005 inch^2 or 0.013mm^2) cell sizes needed for the 65-kilobit chips. The innovations in technology and circuit design that will be needed to reach the 65k level, such as bit sharing, charge coupling, or multi-level memory planes, will resist quick translation into production devices.

Of the several options which show promise for 65-kilobit cell integration, attention is presently being focused on the charge-coupled device (CCD) technique that Texas Instruments has used to build an experimental RAM cell. Known among insiders as the no-transistor RAM the device stores bits in switchable CCD capacitors implanted under the MOS gate. One reason the CCD approach is so attractive is that it lends itself to the same silicon-gate process used in 4k RAMs. The CCD memory cell, which meets the necessary size requirements for 65k integration, can switch as fast as a conventional MOS transistor so no speed is lost. More significantly, the CCD RAM requires only two access lines per cell: one for storage and a sense line for reading. This further reduces chip size (all of today's RAMs need three lines per cell). The question that remains is whether the CCD memory cell can be manufactured in large chips with high yields – if it can't, designers will have to come up with other designs or further develop semiconductor technology.

Although the new 16k random-access memories received a good deal of attention at this year's Solid State Circuits Conference in Philadelphia, a number of other new developments were described which will have great impact on future electronic circuitry. Among the new circuits are Fairchild's new I^2L RAM which puts 4096 bits of bipolar memory on a single chip, a 16-bit minicomputer controller on a single chip from Toshiba, Intel's n-channel static RAM which breaks the 100-ns speed barrier, and a 4k static RAM from American Micro Devices which operates from 5 volts (a first at that density level). Also described was a continuously-charge-coupled random-access memory (C^3RAM) from Siemens in Germany that shows promise for 65k integration. It all adds up to another exciting year for digital electronics.

Jim Fisk, W1DTY
editor-in-chief

presstop

EXCELLENT AMATEUR RADIO PR has resulted from Amateur Operators' extensive on-going contribution in Guatemala. For example, WB2JSM at the Hall of Science Radio Club in Flushing, N.Y. got coverage in both the New York Daily News and New York Times plus CBS TV nationwide exposure for what ended up as a 'round the clock message handling operation. The Red Cross put out an extremely laudatory press release on the Amateur's role in the disaster, and those Amateurs who called their local papers and broadcast stations found very receptive ears.

LONG AWAITED RACES DOCKET has finally received FCC approval and in effect gives the Radio Amateur Civil Emergency Service to Radio Amateurs. In their Report and Order on Docket 19723 the Commissioners discontinued the requirement for RACES communications plans, FCC certifications and authorizations. It also permits RACES station Licenses to be issued directly to civil defense organizations, and provides for the shared use of all the Amateur frequencies by RACES on a first-come, first-served basis except during emergencies requiring invocation of the President's War Emergency Powers.

Control Operators of RACES stations will have to be licensed Amateurs, and operating privileges for RACES operators will be identical to those of the Amateur Service. Use of RACES stations will be limited to bona fide emergencies plus up to one hour per week of drills and tests.

Existing RACES Stations may continue to operate under existing authorizations and the old rules until their present licenses expire. Any presently licensed RACES station whose license expires less than 18 months after the March 23, 1976 effective date of the Report and Order is also permitted to renew the present license for one additional year.

10 METER REPEATERS will now be permitted according to an Order just released by the FCC. Inputs and outputs for in-band ten-meter machines must lie between 29.5 and 29.7 MHz, and cross banding to ten meters will be tricky since the rule change specifically prohibits repeating the transmissions of an Amateur not authorized to operate on the 28-MHz band.

AMATEUR EXTRA CLASS licensees who wish to have the Extra Class certificate should make a written request to the FCC Field Office at which they took the examination. Requests must include a photocopy of the Extra Class license — any requests that go to Washington or Gettysburg will be returned without action.

OSCAR ARTICLE by K3RXK in February Popular Mechanics is a beautifully presented, outstanding presentation of the Amateur space program. Highly recommended reading.

Too Much Power May be being used by as many as 95% of OSCAR users responding to the AMSAT Newsletter poll. Based on equipment and antennas reported, most users should cut back to avoid exceeding the design ERP levels.

OSCAR Users Should start checking 29510 during stateside passes for current news. That frequency has been suggested as an "AMSAT traffic" frequency, primarily for control stations, but general users would profit by learning of schedule changes and other operating modifications as they occur.

Regular "QRP Nights" on the two OSCARs have been suggested by K1HTV — Rich has worked 13 countries while running 1 watt input and recently QSOed W2BXA with only 0.1W ERP! He'd like user suggestions as to how an on-going low power program could be set up for satellite users — write his Callbook address.

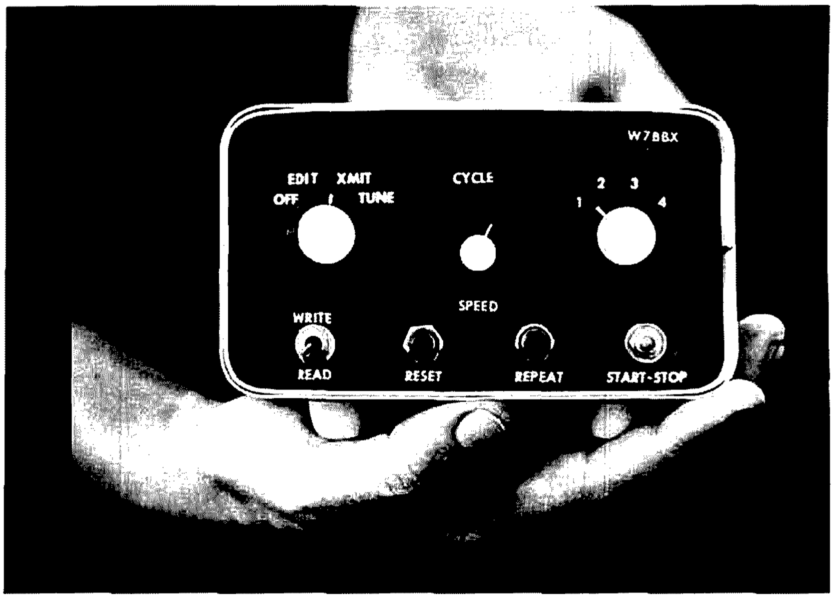
OSCAR 8 Projected Launch period is now starting to shape up, will probably occur sometime in the winter of 1977-1978.

IDENTIFYING YOUR GEAR now increasingly important. For starters, make your markings deep enough that they cannot be easily polished away and place them on the outside where they can be seen without tearing the radio apart. Then use:

1. Your Callsign: not much help to police, but best mark of all at a hamfest — particularly one at which you and your loss are known; plus,
2. The rig serial number: too many radios have easily removed serial plates or tags; plus,
3. Your drivers license number with state as "(NH)ABC-XXX-1234" — this gives police anywhere a path back to you for your recovered gear; plus (optional),
4. Your telephone number with area code — another good way for police tracing, but some don't like it because the thief might also be able to follow it; or,
5. Your name and address — easy to follow by police but just as easily by the thief.

Selling Or Buying — equipment so marked should not be a problem provided both buyer and seller get and keep copies of a descriptive bill of sale. Buyers should also insist that the seller verify his identity, since you're the one stuck if the rig is hot and you can't get back to him.

TWO MAJOR DX CONVENTIONS have settled on September meeting dates — DXPO 76 is set for September 25 in Reston, Virginia, and W9-DXCC will be September 11 in Chicago.



programmable contest keyer

A CW man's keyer featuring high memory capacity, operating convenience, and reasonable cost

A programmable memory keyer is a desirable asset in contest work. It can handle much of the repetitive work while you check dupes, fill out the log, or just take a break. The few programmable keyers on the market all have some desirable features, but they lack the capacity and automatic memory control necessary for smooth, high-speed contest operating. A programmable memory keyer is also needed that the average amateur can afford. The keyer described here has been designed to meet these needs.

Major design objectives included high memory capacity, low cost, and operating simplicity for both program and readout modes; manual, semiautomatic, or fully automatic operation; nonvolatile, nondestructive memory readout; and convenient size. The design is centered around the Intel P2102, a 1024-bit static programmable random access memory (PRAM) in a 16-pin package.* This IC was selected because it requires no refresh circuitry as do dynamic PRAMS, only a single +5 volt power supply is required, all inputs and outputs are fully TTL compatible, and it's readily available at reasonable cost.

description

The keyer (**fig. 1**) is designed so that manual operation with a paddle or bug will always override the mem-

*Intel, 3065 Bowers Avenue, Santa Clara, California 95051.

By **Howard F. Batie, W7BBX**, 12002 Chevoit Drive, Herndon, Virginia 22070

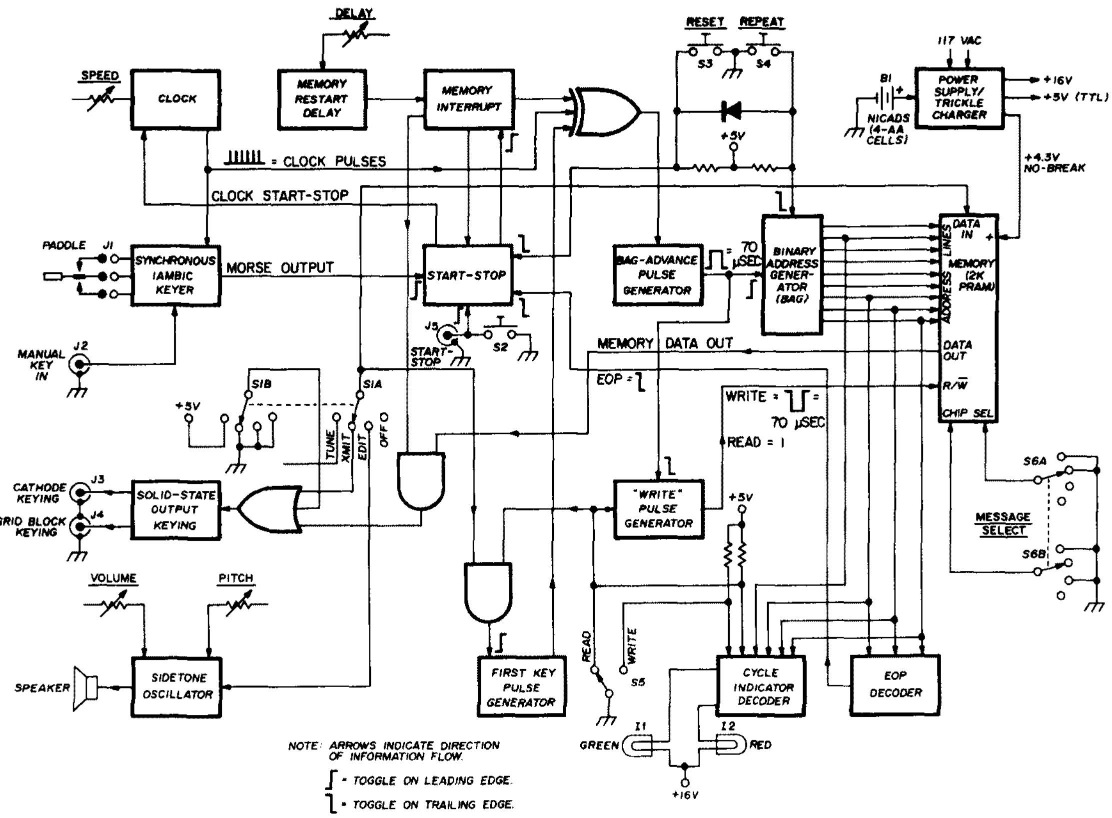


fig. 1. Simplified block diagram of the programmable contest keyer designedby W7BBX. Features include iambic keying, four selectable 512-bit memories, built-in sidetone oscillator, and solid-state transmitter keying.

ory readout. Operation is identical to that of a conventional digital iambic keyer when the memory section isn't used. The popular clock and iambic keyer described by Garrett[1] were modified slightly to interface with the memory. The synchronous clock begins at the instant the paddle is closed and runs for two clock pulses after character generation ceases. The self-completing characters are perfectly formed and spaced throughout the speed range, and character generation is jam-proof. Speed is continuously and smoothly variable from about 8 wpm to well above 60 wpm. The dot memory allows automatic insertion of a dot while holding the dash paddle closed. Similarly, a dash may be inserted while holding the dot paddle closed. Iambic operation allows alternate perfectly spaced dots and dashes to be generated when both paddle arms are simultaneously closed. An external manual key or bug can be used directly instead of the paddle and will control all keyer and memory readout functions.

Solid-state output keying for all inputs (paddle, external manual key and memory readout) is incorporated. The keyed output is directly compatible with most popular cathode-keyed and sidetone/vox actuated and grid-block-keyed transmitters; two output keying jacks, one for positive-keyed voltages up to +150 volts and one for negative-keyed voltages up to −150 volts, are simultaneously available on the rear panel. However, if your transmitter is cathode keyed and 100 mA or more flows through the keyed circuit, an external pass transistor or keying relay may be required. A twin-T audio oscillator and amplifier provide a sinusoidal sidetone waveform that drives an internal 8-ohm permanent-magnet speaker with sufficient audio to perform well in a moderate ambient noise environment. Volume is adjustable, and the pitch is variable from about 400 to 1500 Hz. The internal sidetone oscillator is activated only during the edit mode; that is, for off-the-air programming or checkout of a programmed message. During transmit, or while programming on the air, the transmitter sidetone oscillator would be used in the usual manner. If your transmitter doesn't have an internal sidetone oscillator, a minor wiring modification to the function switch S1A terminals will permit the keyer's internal sidetone oscillator to be used in both edit and transmit modes. A *tune* position is incorporated for tune-up purposes.

memory readout

With S5 in *readout* (fig. 1) and the stored message to be transmitted selected by S6, readout is initiated by depressing S2. This starts the clock, and the clock pulses are fed to the binary address generator (BAG), which includes nine tandem flip-flops. As the flip-flops cycle

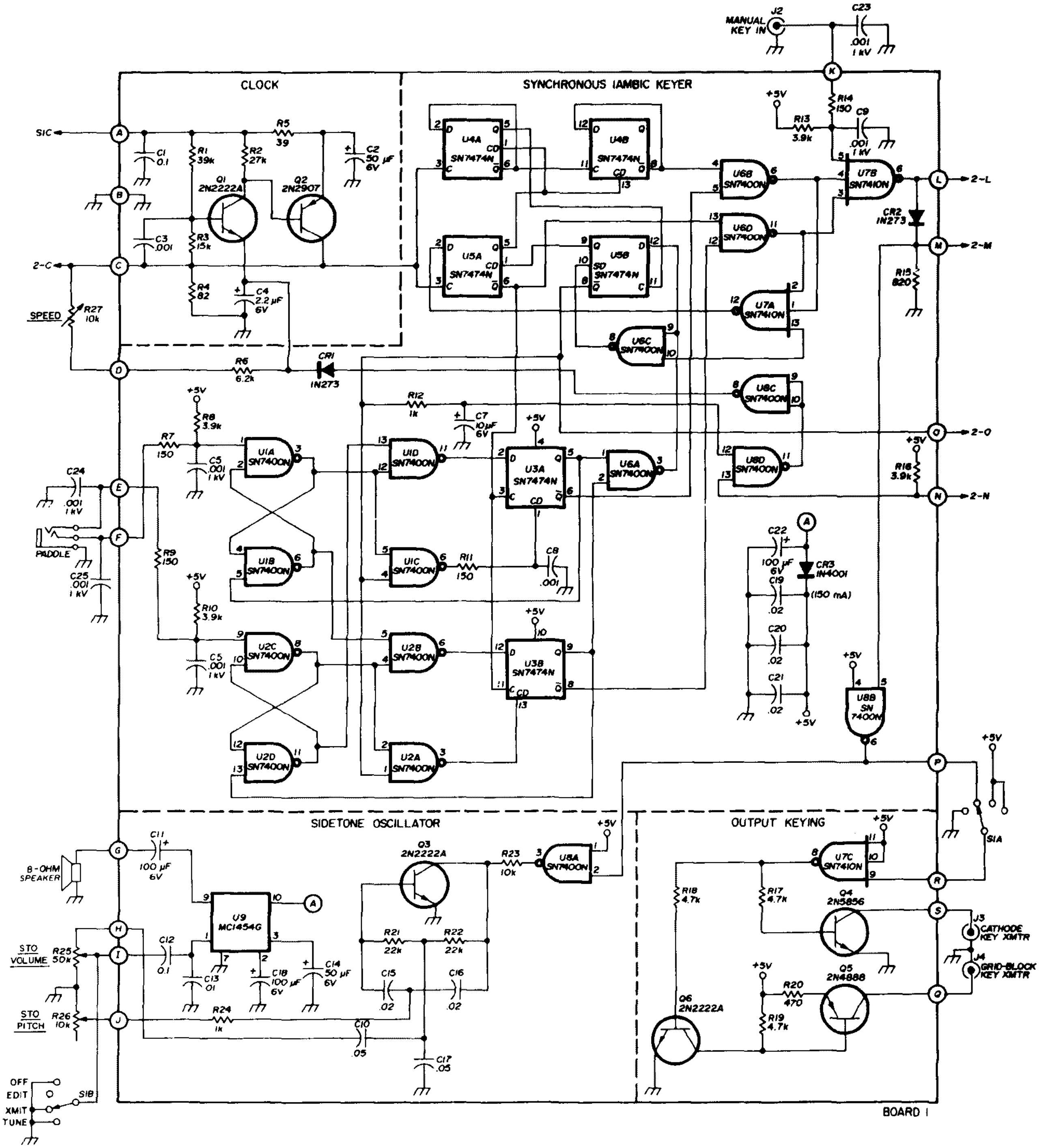
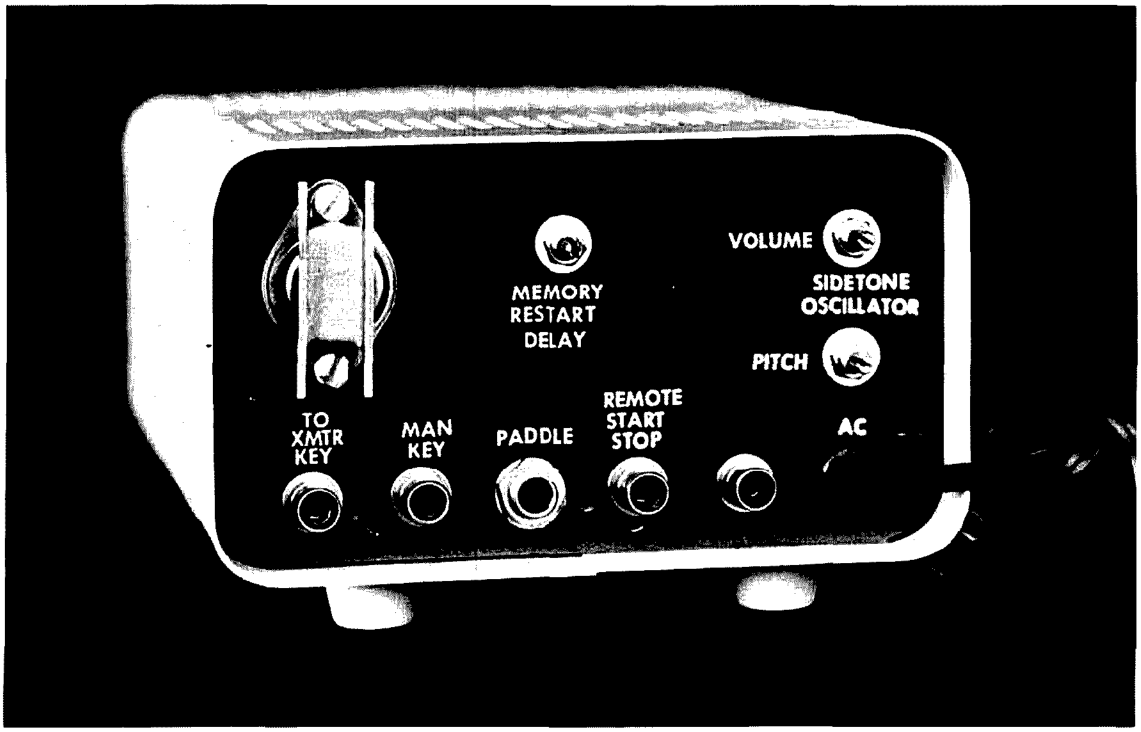


fig. 2. Logic diagram of the keyer board for the programmable contest keyer. Designations 2-C, 2-L, etc., indicate connections to board 2 (fig. 3). S1 is a 3-pole, 4-throw shorting-type rotary switch. All resistors are ¼ watt, 10%.

through 511 successive counts, their BCD output is applied to the nine address lines of the selected memory section, and the addressed information stored in each memory cell is automatically presented to the memory chip data out terminal. If desired, memory readout can be halted in mid message by depressing S2 again before completion of the entire readout sequence. Further memory readout is inhibited until S2 is again depressed;

memory readout will then continue from the point at which it was interrupted (semiautomatic operation). On the 511th clock pulse fed to the binary address generator, the BAG returns to all zeroes on the nine output erator. Depressing S4 during the first seven-eighths of the message readout sequence resets only the binary address generator to the message beginning, which is then automatically repeated. Thus, if "CQ TEST DE W7BBX/4"



Controls and receptacles on rear panel. Although not labelled, one jack is for grid-block keying; the jack labelled "to xmtr key" is for cathode-keyed transmitters.

lines (end of program readout). The downward transition of the highest significant memory address line is unique and signifies "end of program," or EOP. This EOP transition automatically stops the clock, and all control circuitry is simultaneously reset to begin another readout sequence when S2 is next depressed.

Rear-panel provisions are made for remotely starting the memory readout sequence. A separate spst switch in parallel with S2 at J5 can control both *start* and *stop* functions; for example, a simple foot switch can be used to free your hands for the paddle or logging. Alternatively, any external circuit that provides a negative-going TTL-compatible pulse can trigger *readout*. (One possible application might be synchronization to WWV for moonbounce, meteor scatter, or satellite relay operations.)

When the memory readout cycle is initiated, the green *cycle* indicator (I1) lights continuously until 87.5% of the memory contents have been read out; at which point it begins to flash to indicate "nearing end of program." When the message has been completely read out, the green light extinguishes. Depressing S3 at any point in the message *readout* cycle stops the clock and resets all control functions and the binary address generator.

were programmed into the memory, selective repeats by S4 can modify the transmitted message to, for example, "CQ CQ TEST CQ TEST DE W7BBX/4." The increased memory capacity over that of many presently available keyers allows a message length up to that of "The quick brown fox jumped over the lazy dogs back" to be programmed into each of the four separate memories.

An essential feature of a contest keyer is the ability of the paddle to override the memory readout to insert exchange number and/or signal reports in the middle of a programmed message (fully automatic operation). The memory interrupt feature allows you to manually break into any point of the memory readout cycle merely by activating either the paddle or external manual key during memory readout; memory readout is instantly interrupted and remains interrupted as long as manual keying continues. When manual keying stops, an adjustable 1-second delay is introduced by the memory restart delay before the keyer automatically allows memory readout to continue from the point at which it had been interrupted. Memory restart does not have to be manually commanded. Thus, a programmed contest message of "DE W7BBX/4 NR 599 VA BK" can be sent correctly by manually inserting the contest exchange number be-

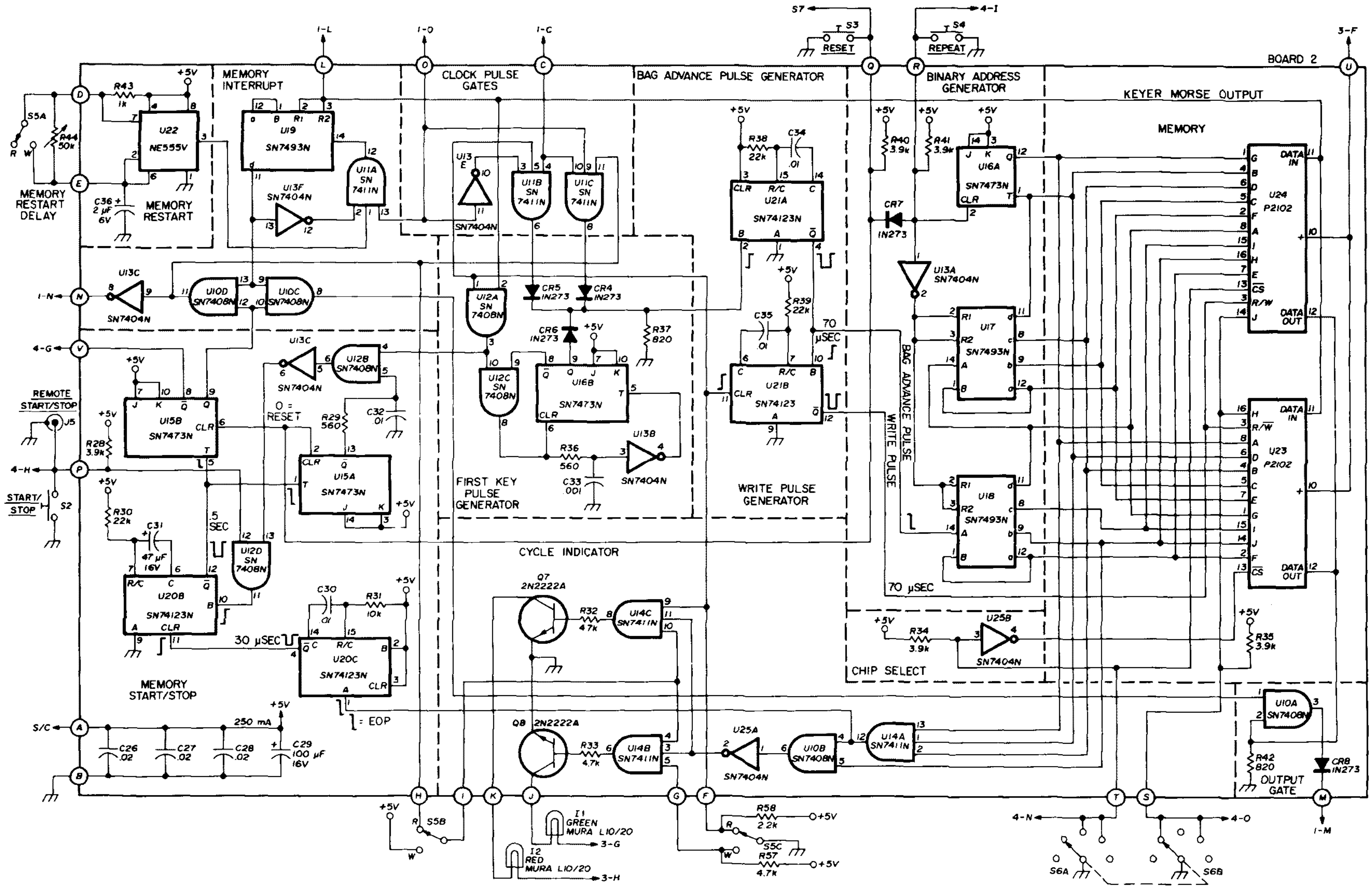
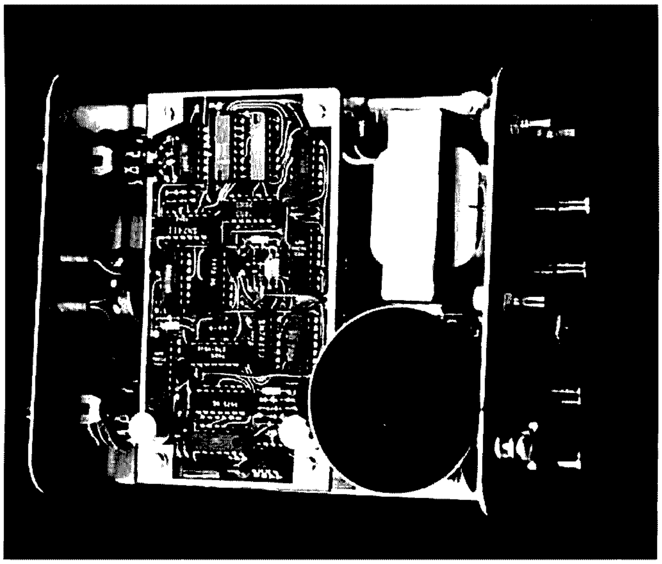


fig. 3. Memory circuit for the programmable contest keyer. Designations 1-L, 1-O, etc., designate connections to board 1 (fig. 2.) Connections 3-F, 3-G go to the power supply, and 4-N, 4-O go to the remote control unit. S2, S3 and S4 are spst push buttons. S5 is a 3pdt toggle switch; S6 is a 2-pole, 5-throw rotary switch (shorting or non-shorting okay). All resistors are ¼ watt, 10%.

tween NR and 599 during memory readout. Memory contents previously stored in the array are automatically prevented from being inadvertently transmitted while the memory is in a hold condition during manual keying.

memory programming

Placing S5 in the *write* position automatically programs a logic zero in the first cell, steps the binary



Chassis top view showing memory board, sidetone oscillator speaker, and power supply transformer.

address generator to the second cell, and causes the red *cycle* indicator (I2) to light immediately, even though the clock is not yet running and nothing is being written into the memory register. The clock is started and programming begins automatically merely by activating the paddle. During the *write* sequence, the clock operates in a "semi-synchronous" mode: while keying normally with the paddle, operation is fully synchronous; if character generation ceases, the clock continues to run asynchronously through the remainder of the message capacity, and logic zeroes are programmed to erase any previously stored message.

The red *cycle* indicator begins blinking when 87.5% of the memory has been programmed and returns to steady red at the end of the programmable capacity; this reminds you to place S5 to the *read* position before initiating a *readout* sequence with S2, or again activating the paddle before a new memory register is selected by S6. Otherwise, the message contents just programmed might be erased.

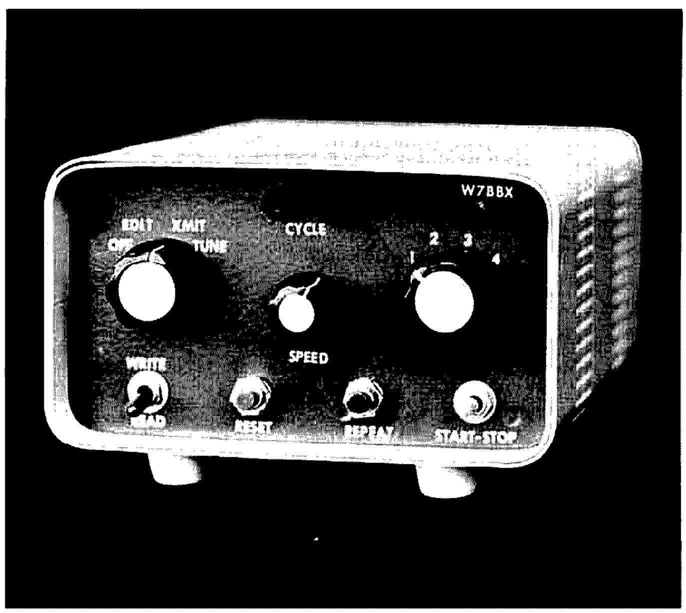
With S5 in the *write* position, the *write* pulse generator is activated. The binary address generator "advance" pulses toggle the BAG on the leading edge of each positive-going pulse, while each trailing (falling) edge triggers the *write* pulse generator to provide the negative-going *write* command to the memory array. Thus, correct timing occurs for accurate memory cell selection and for writing into the selected cell the logic level that appears on the data-in line (keyer output) at the instant of the *write* pulse. S1 and S5 are independent, so the keyer may be programmed on the air (S1 in *transmit* mode) or off the air (S1 in *edit* mode).

power supply

Although the Intel P2102 has nondestructive memory readout (stored information is not lost during readout), loss of power to the memory chip causes loss of the entire stored information (the memory chip is volatile). To keep Murphy and his despicable laws out of the memory, a no-break trickle-charged nicad supply is recommended. Such a charger will preserve keyer memory contents for about 2 to 3 hours, which will eliminate reprogramming when your Field-Day generator runs out of gas. Completely discharged nicads will be recharged in about 20 hours.

remote operating control

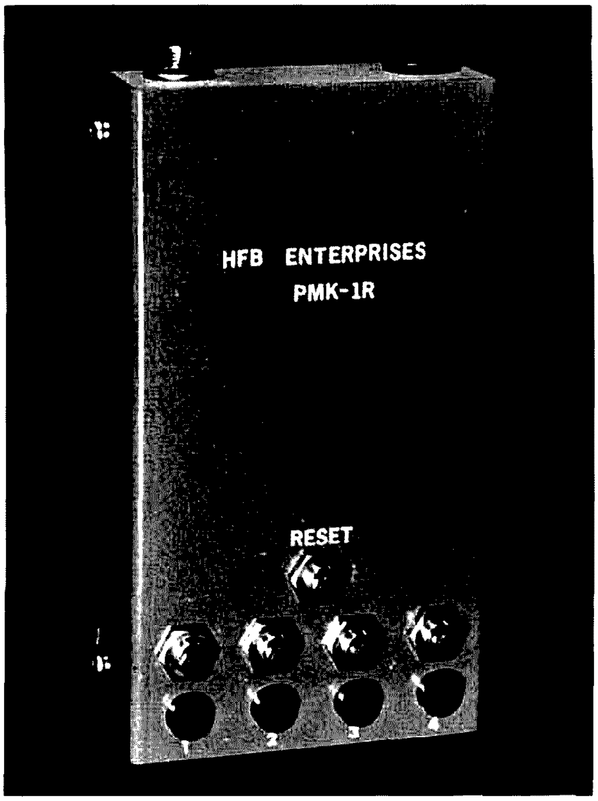
Provisions can be made on the rear panel of the contest keyer to accommodate a remote operating control which can be conveniently placed next to your paddle or bug.* The remote unit controls those keyer readout functions which are most necessary during a contest: message selection, message start, message repeat, and message reset. Depressing any one of the four message select pushbuttons antomatically selects that message, resets the memory to the message beginning and starts meassage readout. Since message selection is



Front panel of programmable contest keyer. Set and forget controls, and input and output jacks, are on rear panel.

*Schematic diagrams for the power supply and remote-control unit will be sent to interested readers upon receipt of a stamped, self-addressed envelope.

independent of the last message sent, successive selections of the same message immediately repeats that message from its beginning. A separate *reset* pushbutton is included to immediately stop the readout sequence.



Remote control unit for the programmable contest keyer provides control of all major keyer functions.

The remote cabinet selected (LMB CR-531) is approximately 1x3x5 inches (2.5x7.5x1.3cm) and houses the five pushbuttons, the printed-circuit board and four optional panel lights which indicate the selected message. Connection to the keyer is made by a plug-in, shielded, 8-conductor cable. All power for the remote is derived from the keyer, and removal pf the remote cable from the keyer does not affect keyer operation.

construction

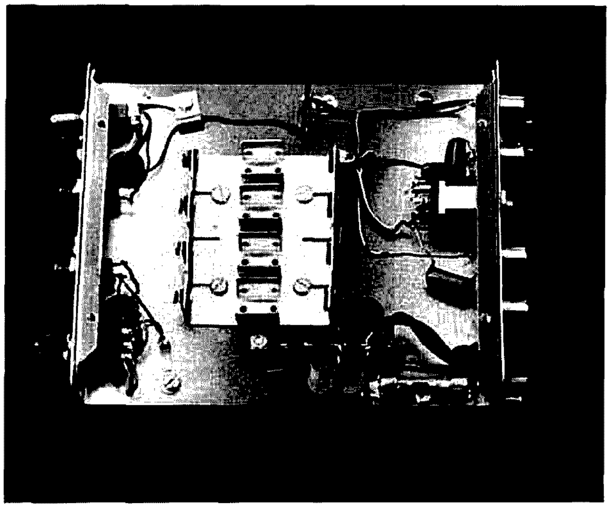
Panel clutter was avoided by automating as many memory-control functions as possible. Most-used controls are on the front panel; others are mounted on the rear of the keyer. An LMB CO-3 enclosure was used. The circuit is mounted on three PC boards: one for the basic iambic keyer, output keying, and sidetone oscillator; one for all memory functions; and one for the power supply.†

If desired, the keyer may be built without the memory board and used as a conventional iambic keyer and the memory may be added later. No keyer-board changes will be required and only four wires need be interconnected between keyer and memory board. The PC-board layout allows all memory-board wires and connections to be added without removing either keyer board or power supply from the cabinet.

A high degree of rf immunity is achieved by a) using TTL instead of CMOS devices, b) providing rf bypassing on all paddle and keying leads, and c) providing a grounding bond between cabinet sections. The keyer has been successfully kilowatt tested at a 3:1 vswr from 80 through 10 meters.

summary

The trade between performance, cost and circuit simplicity is sometimes difficult. For this project it was decided to opt for a capacity of four 512-bit messages and gain the advantages of paddle-programming and fully-synchronous operation at the expense of increased circuit complexity. This decision has been proved by the keyer's flexible, reliable, and unconfusing operation. This keyer has been a most useful operating aid both at home and in the field under generator power (thanks to the nicads). I'd like to thank the members of the Potomac Valley Radio Club for their constructive comments, suggestions and support.



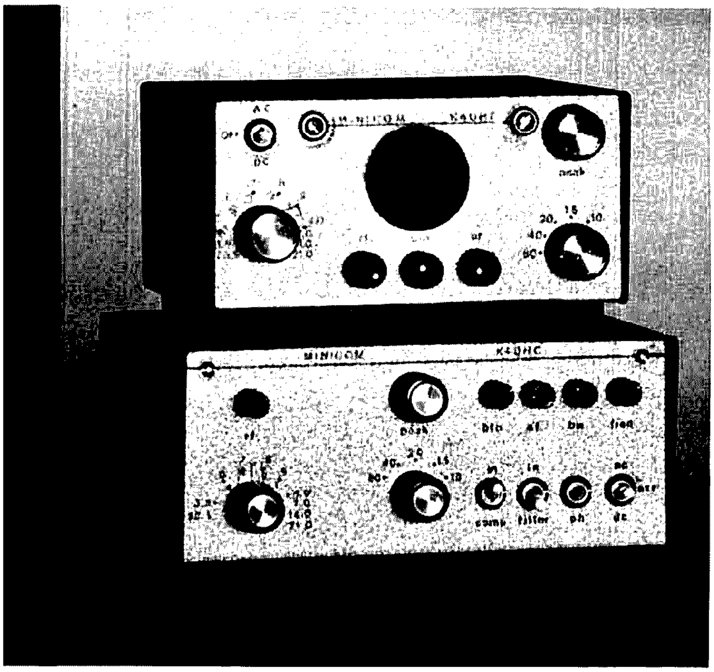
Underchassis view showing battery receptacles and power-supply wiring.

†A set of double-sided PC boards with plated-through holes plus assembly and operating instructions are available from HFB Enterprises, Post Office Box 667, Herndon, Virginia 22020. The price is $30.00 post paid. Included are step-by-step assembly instructions, a complete parts list, operating instructions, and a full set of drill templates for the LMB CO-3 and LMB CR-531 cabinets.

reference

1. James Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.

ham radio



design ideas for miniature communications receivers

A collection of miniaturized receiver circuits including the Minicom Mk III, a five-band converter, an 80-meter tuner and two complete i-f and audio systems

I wrote my first article for an amateur publication in 1971.[1] Writing was something I had always wanted to do but until then I had never gotten up enough courage to try it. After several weeks of writing, rewriting and drawing circuit diagrams, I felt I had something that might be accepted by the editor. When it was finally published I discovered what the real rewards were: letters came in from all over the country and from many DX locations. I not only had a great time answering all these people and swapping ideas with them, I made many new friends with whom I still correspond.

Of all the articles I've written since then, only my most recent one generated as much enthusiastic response as the first. Both of these articles discussed receivers and miniaturization; evidently this is a favorite topic with the majority of readers. If you found the Minicom[2] interesting, read on.

The original Minicom packed a lot of performance into a small package, but

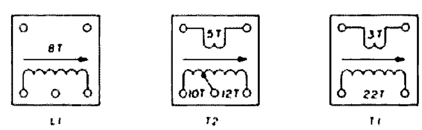


fig. 2. Bottom view of inductor L1 and transformers T1 and T2. All coils are wound with salvaged wire on standard 3/8" (10mm) i-f transformers.

even before the article had been published a new version was on my workbench. It packed even more performance into the same space and was called the Mk II. If you were among those who

Ray Megirian, K4DHC, Box 580, Deerfield Beach, Florida 33441

wrote to me regarding parts for the Minicom, you already know about the Mk II as an information package was made available to interested builders. Since then the Mk III has been devised and will be discussed here along with several other useful assemblies for ssb/CW receiver construction.

Minicom Mk III

My particular use for the Minicom has been as a tunable i-f for use with an external converter for all-band coverage up to 30 MHz. Several such converters have been built in very compact form so that, when teamed up with the Minicom assembly, it results in a very diminutive multi-band communications receiver.

Development of the Mk III receiver was undertaken for several reasons. First, I was interested to see if the receiver could be made smaller without resorting to techniques that would be impossible to duplicate. I also wanted to determine the feasibility of substituting varactor tuning for the three-gang tuning capacitor. If successful, the latter would not only eliminate a major hard-to-find component, it would make miniaturization a reality.

Some time ago Motorola announced their MVAM-1 triple tuning diode and it was this item that encouraged me to go ahead with my investigation of an electronically-tuned Minicom. Although designed for tuning a broadcast-band receiver, a study of the MVAM-1 data sheet led me to believe that diode Q

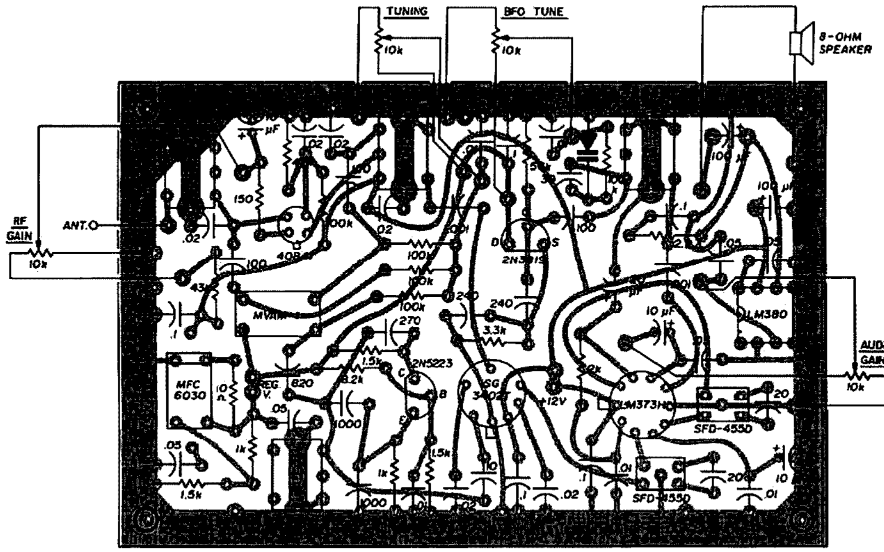
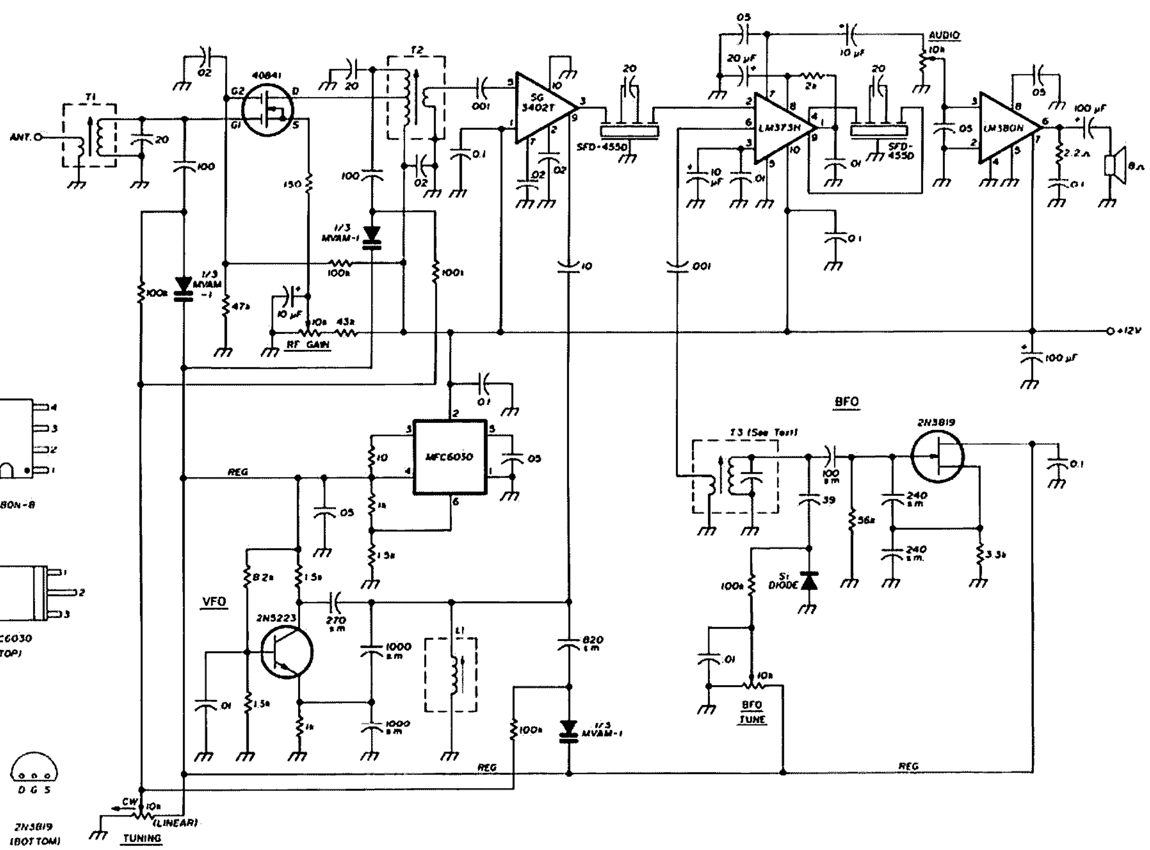
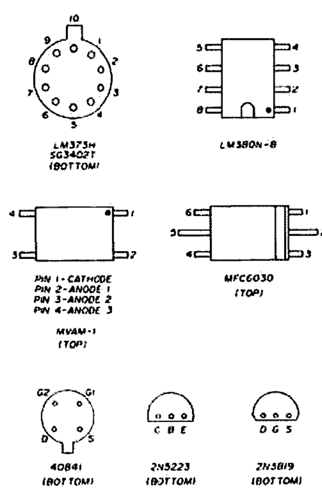
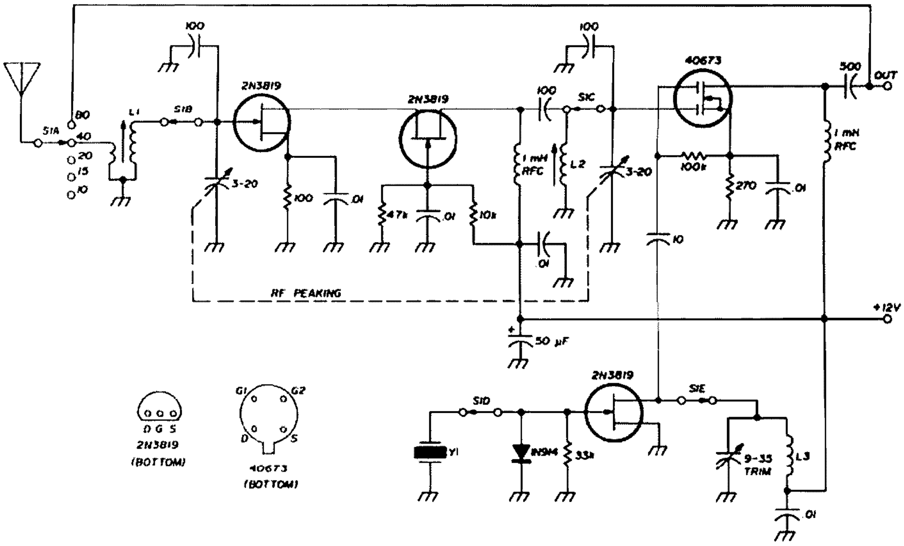


fig. 3. Printed-circuit layout for the Mk III communications receiver.

fig. 1. Circuit diagram of the MkIII communications receiver which uses tuning diodes. Remove pins 4, 6 and 8 from the SG3402T IC before installing it on the circuit board. Details for L1, T1 and T2 are shown in fig. 2. All resistors are ¼ watt, 5%.



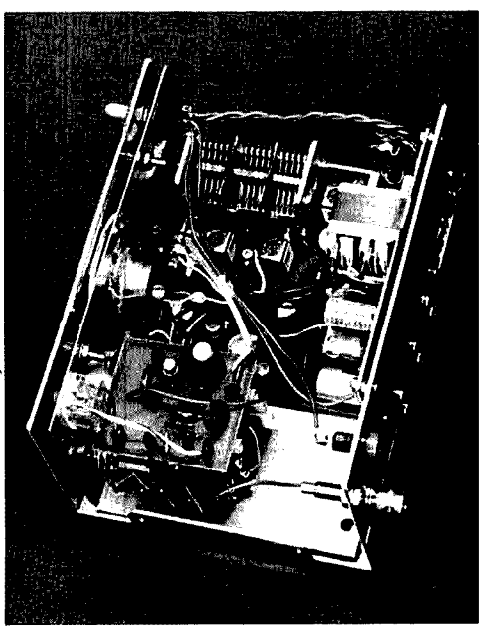




band	L1	L2	L3	Y1
40	20 turns no. 36 (0.13mm), 2 turn link	20 turns no. 36 (0.13mm)	4.7 µH	11 MHz
20	12 turns no. 28 (0.3mm), 1.5 turn link	12 turns no. 28 (0.3mm)	2.2 µH	18 MHz
15	7 turns no. 28 (0.3mm), 1 turn link	7 turns no. 28 (0.3mm)	1.5 µH	25 MHz
10	4 turns no. 28 (0.3mm), 1 turn link	4 turns no. 28 (0.3mm)	1.5 µH	25 MHz

fig. 4. Five-band converter for use with the Mk III communications receiver. In this design the 25-MHz crystal is used for both 10 and 15 meters. Complete coverage of the 10-meter band requires additional crystals. Inductors L1 and L2 are wound on 0.215" (5.5mm) diameter PC-mount coil forms with Carbonyl-E (red) cores. Molded rf chokes are used at L3.

Interior of the smaller receiver. The Mk II board assembly is mounted on standoffs to the sub-chassis (top). The converter is in the foreground and the power supply is on the rear panel.

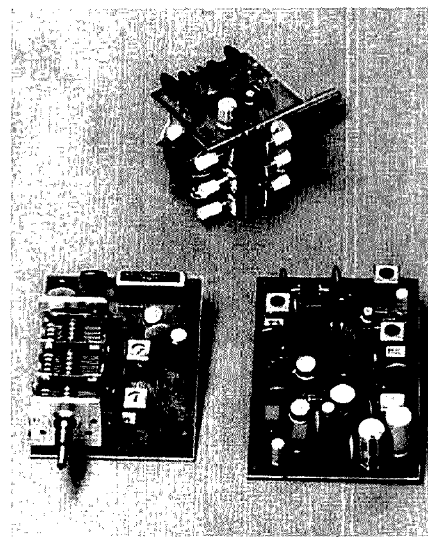


would be high enough at 4 MHz to produce a pretty good receiver. The final results achieved with a breadboard receiver were quite satisfactory and led to the layout shown here. The board is 2.5 inches (65mm) wide by 3-7/8 inches (100mm) long. The Mk II is 3.75 inches (95mm) square. The schematic of fig. 1 shows the circuitry for the receiver which tunes from 3.5 to 4.0 MHz using approximately 80 percent of the tuning pot rotation. Tracking is good within these limits but starts to deteriorate at the extreme ends of the resistance element.

One big advantage of the diode tuning system is mounting flexibility. Since tuning is accomplished with a potentiometer at the end of a three-wire cable, the receiver may be mounted anywhere in the cabinet without worrying about exact positioning or shaft alignment.

To make tuning easier, some sort of reduction drive is needed, just as with conventional tuning systems. Perhaps the reduction drives that are the easiest to find are the small Japanese vernier dials which have better than 7:1 reduction. The built-in stops which limit travel to 180 degrees can be eliminated by snipping off the stop pin, thus allowing continuous rotation.

Except for the mixer, which uses a Silicon General SG3402T IC, the circuitry of the Mk III is quite conventional. Good results with this mixer were achieved in the Mk II so it was used in the Mk III as well. The LM373H once again does a commendable job in the i-f section where two Murata SFD-455D ceramic filters provide selectivity. An LM380 replaces the MC1454 in the audio portion, adding the advantages of electrical and thermal overload protection. The bfo remains unchanged.



The 40-, 20-, 15- and 10-meter converter is at the top. The electronically-tuned Minicom Mk III is on the right; on the left is the 80-meter tuner.

diode tuning

A few words about the MVAM-1 might be worthwhile at this point since that's what the Mk III receiver is all about. The three diodes are wired with a common cathode lead and housed in a tiny four-pin plastic package. Typical capacitance at 1.0 volt reverse bias (V_R) and f = 1.0 MHz is 480 pF. The capacitance ratio is 15 (minimum) for reverse bias between 1 and 25 Vdc. The three diodes are matched within ±1.5 percent over the entire capacitance vs voltage

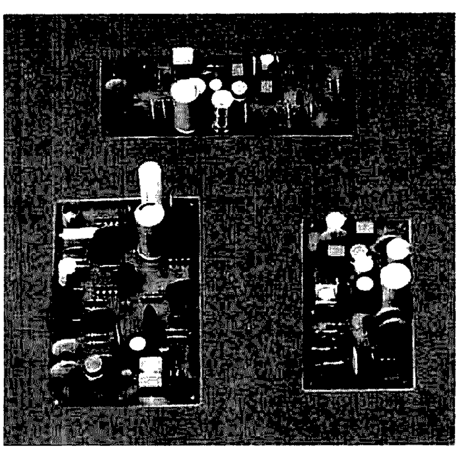
curve. Typical Q at V_R = 1.0 Vdc and f = 1.0 MHz is more than 500.

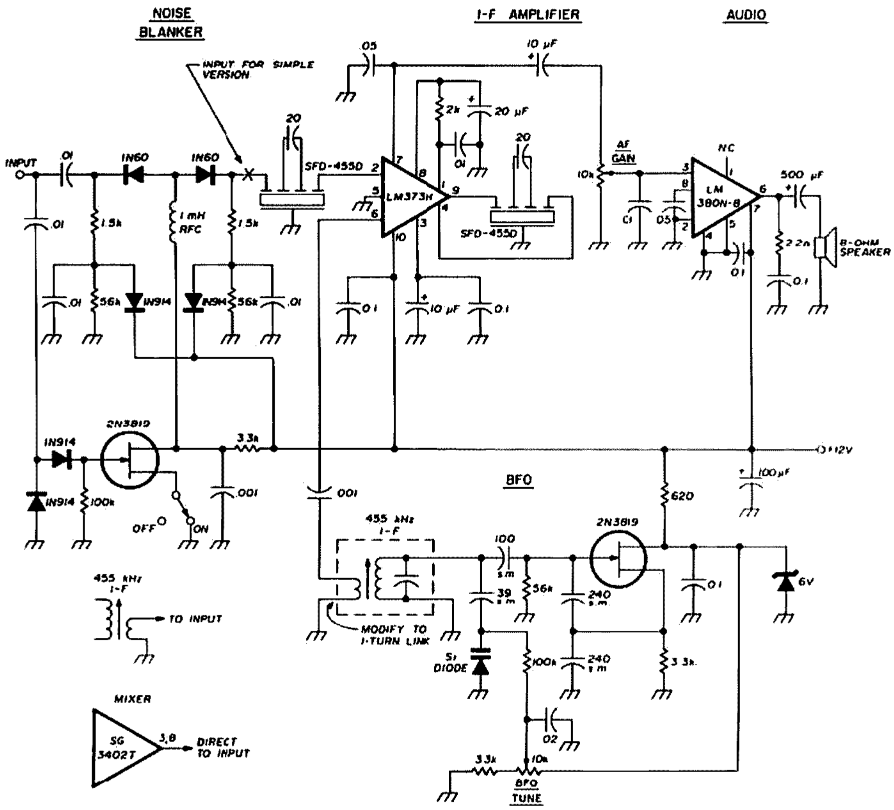
Because of the large capacitance rating of these diodes, it was possible to change the vfo configuration to a more drift-free design. This new vfo circuit drifts very little; so long as the receiver is operated in a relatively stable temperature environment, drift is practically nonexistent. The diode capacitance temperature coefficient could be a problem, however, if the receiver were operated under widely varying ambient temperature conditions. The diode capacitance temperature coefficient given in the data sheet is typically 435 ppm/°C at V_R = 1.0 Vdc and f = 1.0 MHz.

Another consideration requiring strict attention with diode tuning is voltage regulation. It stands to reason that since the diodes are voltage-controlled devices, the source of control voltage must retain a high degree of stability. A simple zener diode is far from satisfactory in this application so an IC voltage regulator (MFC6030) was included on the circuit board to provide the needed regulation. With the values shown in fig. 1 the output is around 7 volts and the minimum input voltage to maintain regulation is 10 volts. Spare terminals on the circuit board allow access to this regulated voltage for use with other external circuits, if needed. On the board the regulator supplies power to both the vfo and bfo besides providing a stable control voltage for tuning purposes.

Standard 455-kHz transistor i-f trans-

The three versions of the i-f amplifier. With noise blanker (top), without (right) and compressor/filter version (left).





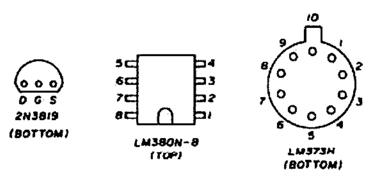


fig. 5. I-f and audio amplifier circuit. Two methods are shown for coupling into the amplifier. The noise blanker may be deleted, if desired. Printed-circuit layouts are shown in fig. 6 (with noise blanker) and fig. 7 (without noise blanker). All resistors are ¼ watt, 5%.

formers were stripped and used for coil forms as in all previous models of the Minicom receiver. If you've forgotten how to strip these devices, refer to the original Minicom article for details.[2] Any of the transformers (white, yellow or black core) may be used for the bfo transformer after one simple modification. Unsolder the secondary (link) leads from the base pins and gently break off the wires where they disappear into the bobbin. Substitute a single-turn link wound over the existing coils using a piece of salvaged wire from the rf coils (see fig. 2). This change assures proper bfo injection level for the LM373.

multi-band converter

A suitable mechanical design for the converter proved to be the most difficult problem of all to solve. After numerous attempts it looked as if the converter would occupy as much space as the Minicom; this put a big dent in my miniaturization program. I finally solved the problem by using three separate circuit boards instead of one. The heart of the assembly is the bandswitch, around which everything else is built. The miniature bandswitch was a surplus item originally manufactured by Oak and has seven 1-inch (25mm) wafers, six of which were single pole, eight position devices. The seventh wafer had a weird switching arrangement which was of no use. Since only five poles were needed, this last deck plus one of the others were removed. The added space was taken up by two of the circuit boards and some suitable spacers. A shield was also installed between sections.

Unfortunately, these surplus switches were a one-time deal and I don't know

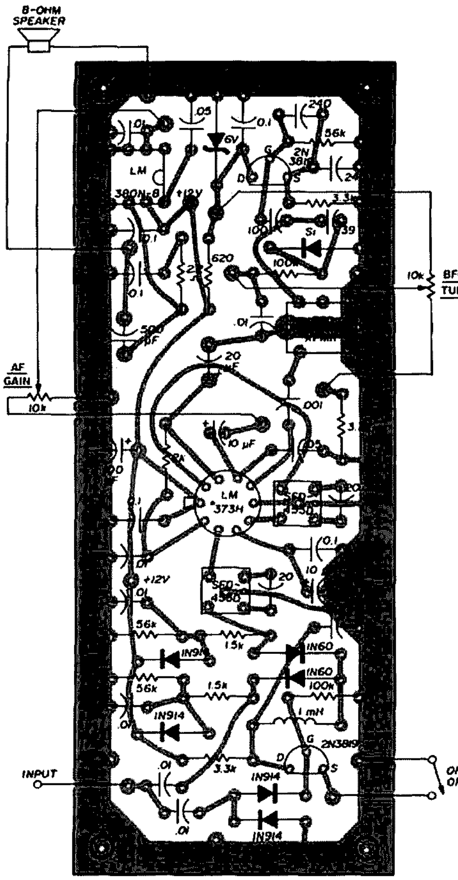


fig. 6. Printed-circuit component layout for the i-f and audio amplifier with noise blanker.

where you can get any of them now, so any detailed assembly instructions for the converter would be useless. However, if you have the skill required to build any of this equipment, you shouldn't find it too difficult to come up with a layout to fit your own components. A schematic of the converter I used is shown in fig. 4.

In my converter only the 28.5 to 29.0 MHz segment of 10 meters is covered. For additional coverage, you'll need additional crystals. The rf section should require no changes to cover the entire 10-meter band. The two-gang tuning capacitor used to peak the converter front end is a tiny film dielectric type removed from a transistor fm radio. If you use one of these, a shaft has to be made.

i-f and audio

Occasionally the need arises for an i-f system which you can use for testing new tuners or front ends. At other times this *sectionalized* construction technique is the only one which will allow cramming everything into the available space. In either case, one of these i-f modules may be just what you need.

An inspection of the schematic in fig. 5 reveals that the circuit is identical to that used in the second half of the MK III. An LM373H IC with two Murata SFD-455D ceramic filters fulfills the requirements for i-f amplifier, detector and agc functions. The bfo is identical to the Mk III as is the LM380 used for the audio amplifier.

If desired, a noise blanker may be included ahead of the i-f input by making the circuit board slightly longer. In either case, the board is 1.75 inches (45mm) wide with a length of 3-3/8 inches (86mm) for the simple version and 4-5/8 inches (117mm) with the noise blanker.

The circuit for the noise blanker is similar to one used in a commercial Japanese two-meter ssb transceiver. A reader in West Germany, Earl Lagergren, sent me the circuit after reading my article on a solid-state noise blanker.[3] Since this noise blanker circuit requires no special components and works at various i-f frequencies, I've included *it* here as a worthwhile addition to your potpourri of receiver circuits.

One of the complete receivers shown in the photos uses this tuner, the four-band *converter* and an *i-f* and audio system (described later). It is identical to the rf and mixer stages of the Mk II which some of you are familiar with.

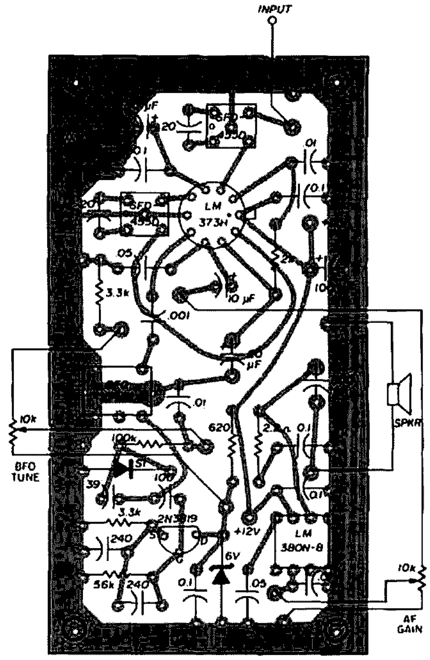
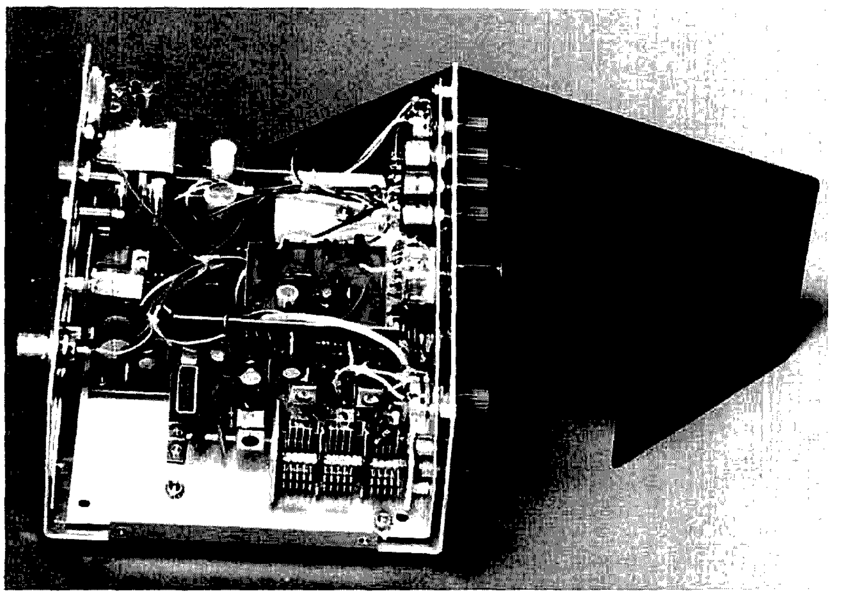


fig. 7. i-f and audio amplifier printed-circuit layout. Circuit board which includes a noise blanker is shown in fig. 6.



Interior of the larger receiver. The tuner was mounted on edge with the converter right behind it. The i-f, compressor/filter assembly is mounted on the sub chassis to the rear. A power supply is installed on the rear panel.

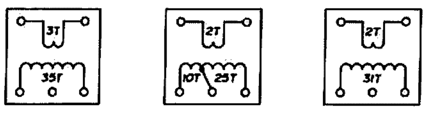


fig. 9. Bottom view of transformers T1, T2 and T3 used in the 80-meter tuner. All coils are wound with salvaged wire on standard 3/8" (10mm) i-f transformers.

Included on the tuner board is a Murata CFS-455J ceramic ladder filter which provides 3 kHz bandwidth at 6 dB down and superior skirt selectivity to systems using the SFD-455D filters.

The rf stage shown in fig. 8 incorporates one of the newer types of mosfet, a dual-gate n-channel enhancement mode, Signetics type SD304. These transistors operate with all-positive bias which may be of interest to anyone who has been looking for an agc system not requiring a dual polarity power supply. About 40 dB of agc range is possible with 0-6 volts applied to gate 2. This circuit, however, uses a manual rf gain control.

The three-gang tuning capacitor is the same as that used in the original Minicom receiver. The two 22 pF padding capacitors for the rf and mixer tank circuits are mounted directly on the variable capacitor before it is installed on the board. These two sections also have built-in mica trimmers for tracking adjustment.

An extra stage of i-f was included on the tuner board to overcome the insertion loss of the ladder filter. A SG3402T IC is used in the mixer section; pin 6 must be removed before mounting the IC on the circuit board.

i-f and audio system

Some time ago I wrote an article describing a complete audio system for use in a communications receiver.[4] This system consisted of a tunable filter, an audio agc system and a power output stage and drew pretty good response from readers. In the version shown in fig. 11 the i-f stages have been included along with the compressor and tunable audio filter. The notch function, however, has been deleted and the circuit has been modified for single supply operation.

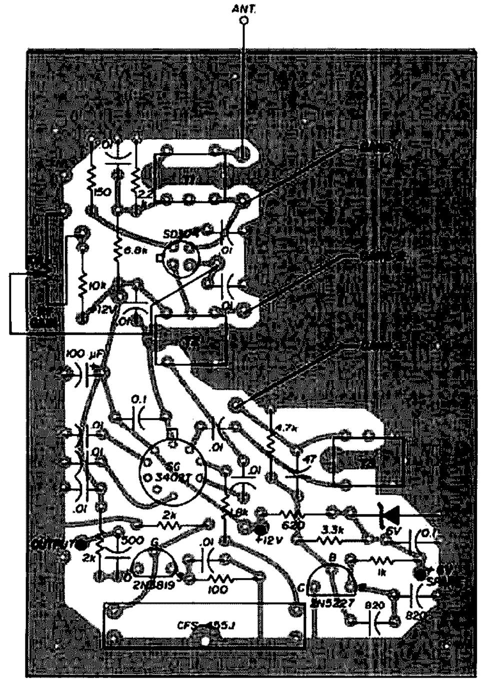


fig. 10. Printed-circuit layout for the 80-meter tuner.

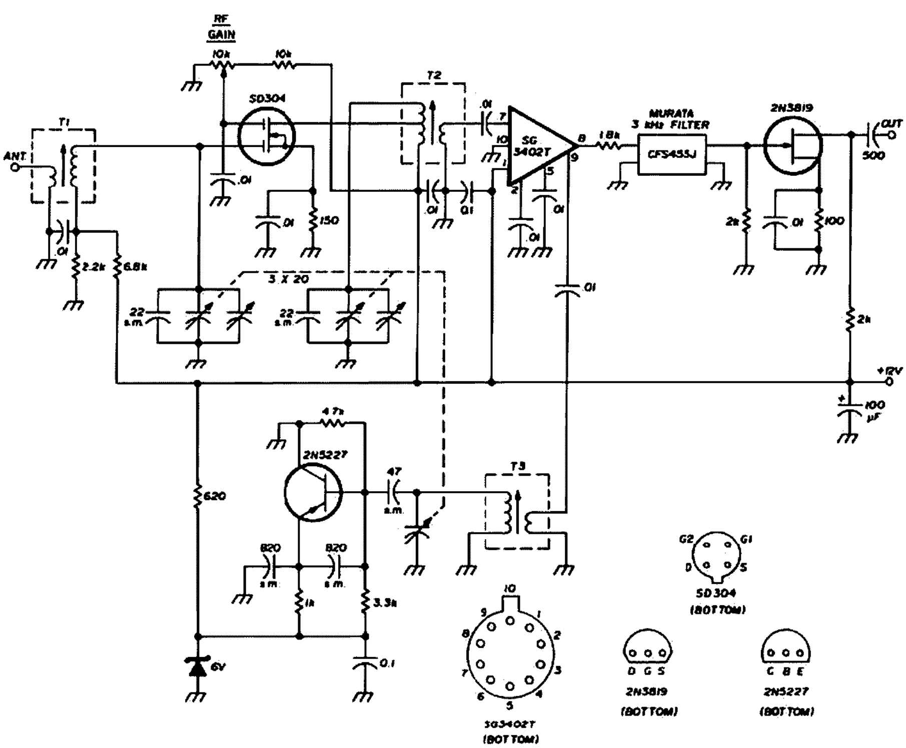


fig. 8. Circuit for an 80-meter tuner which incorporates a ceramic ladder filter. An additional i-f stage is provided to compensate for the insertion loss of the filter. Remove pin 6 on the SG3402T IC before installing it on the PC board. Transformer construction is illustrated in fig. 9. Printed-circuit layout is shown in fig. 10. All resistors are ¼ watt, 5%.

In this circuit two op amps are used in the compressor and two more in the filter. Dual type N5558V units were used and any of the equivalents from other manufacturers may be substituted. An LM380 is once again used for the audio output stage. The i-f portion is the same as all previous LM373 configurations except for the SFD-455D filters. Since I used this assembly in conjunction with the tuner which incorporates the ladder filter, the other units were not required.

The trimmer resistor in the audio filter may be very simply adjusted for proper operation. First set the trimmer for maximum resistance by turning the adjusting screw counter-clockwise. Then turn the *frequency* control to the high end and set *bandwidth* to sharp. Hold your ears, turn on the power and flip the filter switch to *in*. The filter should take off with ear-shattering feedback. Start turning the resistance trimmer clockwise until oscillation ceases (add an additional half turn for good measure). Throw the filter switch back and forth a few times to see how stable things are and adjust the trimmer some

more if needed. The tuning range of the filter should cover from about 500 to 2000 Hz.

It should be pointed out that caution must be used when coupling into this assembly. Note that the i-f input goes directly to pin 2 of the LM373. A dc path must be avoided at this point to prevent damaging the LM373. When used with the 80-meter tuner described above, a coupling capacitor is provided on the tuner board to prevent shorting pin 2 to ground.

Printed-circuit layouts are provided here for all modules except the converter. This should make it easy to duplicate any of the assemblies. The main goal of this article, however, was to present the widest selection of proven and reliable circuits as possible in a single package so that the greatest number of readers could make use of at least one of them. There is nothing sacred about any of the circuits so if you feel like sneaking a piece from one and adding it to another, it will give you a chance to practice your skill at laying out new circuit boards.

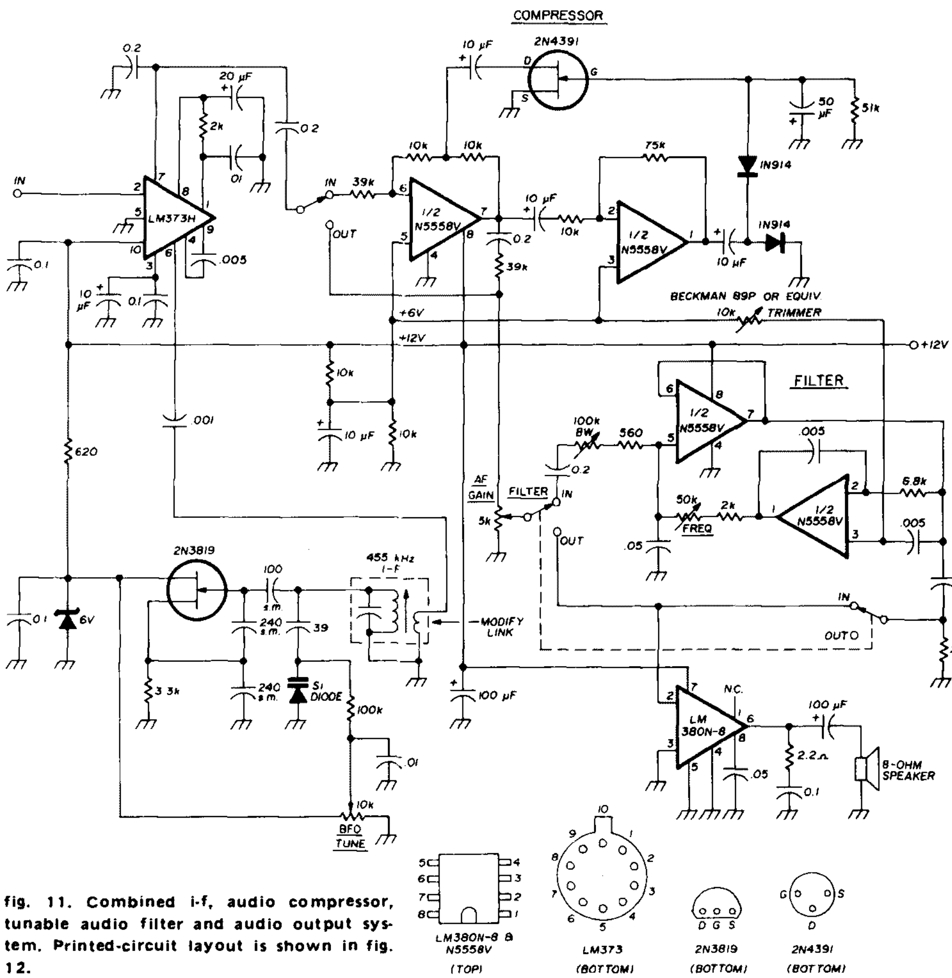


fig. 11. Combined i-f, audio compressor, tunable audio filter and audio output system. Printed-circuit layout is shown in fig. 12.

No tuning or adjustment procedures will be covered since it is assumed that the average builder can accomplish that on his own.

The photos show the various printed-circuit assemblies which have been described plus two receivers built from some of this equipment. The smaller receiver uses a Mk II with the converter described here for five-band coverage. A small ac power supply is attached to the rear panel inside the receiver and a plug is provided for feeding in an external 12-volt dc source. A small speaker is mounted on the front panel to make the unit completely self-contained. The cabinet is 3x5x6 inches (76x127x152mm). The larger receiver uses the tuner, converter and i-f, audio filter and compressor board described above. It also has a built-in ac supply. Cabinet size is 3x6x7 inches (76x152x178mm).

All of the circuits operate from 12-volt supplies and function well down to 9 volts or up to about 14 volts. The Mk III, however, includes a regulator requiring a minimum of 10 volts for proper operation. The copper border around all the circuit boards is the common or supply negative and should be grounded.

With the exception of the board containing the audio filter, the only external components are operating controls. In the case of the audio filter a resistor and capacitor are wired point to point between the front-panel controls.

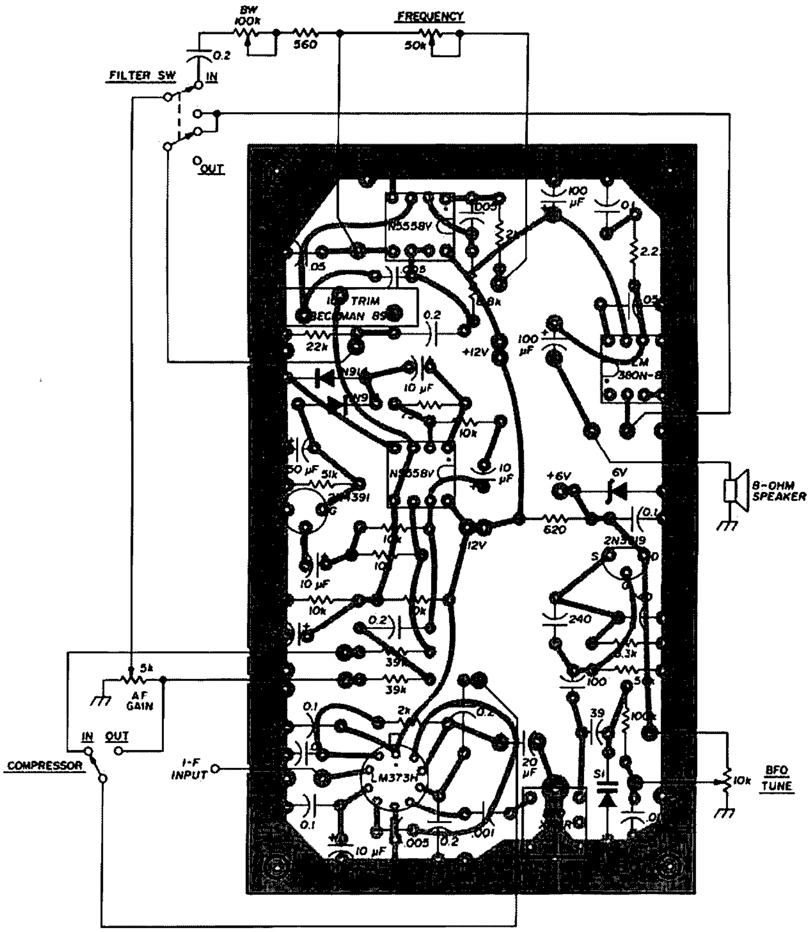


fig. 12. Printed-circuit layout for the combined i-f, audio compressor, audio filter and audio output system.

This saves the extra leads that would be required for going back and forth from the board for these two items.

Before concluding, I should say something regarding parts. Some substitution is possible since there is nothing special about a 2N5227, a 2N3819 or a 2N5223. The 40673 and 40841 can be interchanged but not with the Signetics SD304. The silver mica capacitors should be small case sizes in most instances to fit the PC layouts. The ceramic bypass and coupling capacitors should be small, low-voltage discs. These, as well as 1/4-watt resistors, are readily available and should present no procurement problems.

If you are interested in purchasing some of the more elusive components such as ICs, filters or whatever, drop me a line with a self-addressed, stamped envelope and advise what and how many you need. If there is sufficient demand for particular items, I'll try to obtain what is needed.

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ham radio

missent ID

During intruder watching and observance of amateur signals, I've noticed a considerable number of RTTY transmissions that have ended with an inaccurately sent identification. In some cases, these IDs were repeated in the same incorrect manner on a later transmission. In one occurrence the signature was "W6MTMAI," but W6MTM doesn't have RTTY equipment. I recorded the audio signal on the mark side, then ran off an ink-line tape that clearly showed the five-letter call sign but with more than normal spacing between letters.

It appears that a local station had some rf getting into an IC identifier, causing problems. The operator bypassed the leads to the identification unit, which cleared up the problem. It would be well for anyone using an IC device to send his call and get help to listen to the resulting ID. It should be sent on full power and on every band on which the device is used. Only in that way can one be sure that the call always is sent correctly.

Bill Conklin, K6KA

simple circuit replaces jack patch panel

Shown in fig. 1 is my solution to the ever-present phone jack patch panel used in amateur RTTY stations. I became tired of the conventional panel with cords breaking, so after a little thought, I developed this circuit. The diodes provide the necessary isolation to allow the switch to select the loop desired for a particular piece of equipment.

Dr. Paul Lilly, K4STE

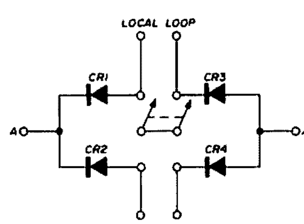

fig. 1. Simple jack patch panel replacement for RTTY stations. Diodes CR1-CR4 provide the necessary isolation, RTTY equipment (printer, keyboard, etc.) is connected to A terminals.

circuit design with the 741 op amp

The inexpensive 741 op amp IC is finding extensive application in electronics — here are some tips for using this versatile device in your own circuits

For several good reasons, the 741 operational amplifier is enjoying wide popularity among amateur circuit designers. It is inexpensive and readily available; several mail-order supply houses are currently selling them for about 35 cents each. The 741 requires no external frequency-compensation networks, and this significantly enhances the ease with which it may be used in circuit development. However, there are several facts about 741s which, if known and understood, can minimize surprises and make their use more predictable.

Those unfamiliar with op-amp theory are referred to a rather extensive article that previously appeared in *ham radio*.[1] That article covers general op-amp circuit design criteria as well as several specific applications. I will attempt here to concentrate on the peculiarities of the 741 and not duplicate material already presented.

The 741 op amp is available in several different packages, but the TO-5 metal can, much like a transistor with eight leads, and the plastic eight-pin minidip package seem to be the most popular. I prefer the minidip because it easily plugs into a dual-inline IC socket. Two 741s can be inserted in a 16-pin IC socket.

Fig. 1 shows the schematic symbol for the 741 op amp with appropriate power supply and null pot circuitry. The null pot is used to set the dc output voltage to zero in certain situations; however, in many circuits

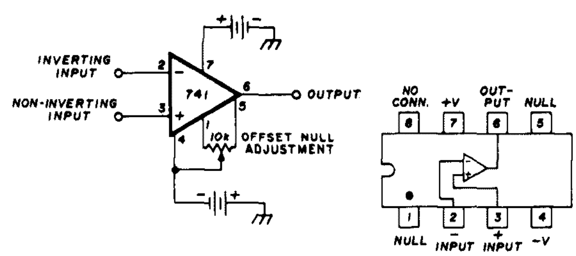

fig. 1. Schematic symbol for the 741 op amp showing power supply and null pot connections. Connections to popular 8-lead minidip package are also shown.

it's not necessary to use a null pot. When the null pot isn't needed, just leave pins 1 and 5 open. **Fig.1** also shows the pin numbering and identification for the 741 minidip when viewed from the top. Either a notch at one end of the package, or a dot in one corner are used to index the pin numbers. The +V and −V pins are for power supply connections.

The maximum rated power supply voltages for the commercial version of the 741 are plus and minus 18 volts, but lower voltages may be used. Two 9-volt batteries will do nicely, but higher supply voltages will permit a larger output signal swing. Current drain will depend on load resistance and output signal amplitude, because the output circuit of the 741 is a class B complementary emitter follower. Under zero-signal conditions,

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive Dallas, Texas

however, the quiescent power supply current should be about 1 mA or less.

dc offset problems

Since the 741 is a dc amplifier, any dc voltage which exists between the two input terminals will be multiplied by the gain of the amplifier and can result in a large dc offset at the amplifier's output. This offset prevents maximum linear output signal swing. In some cases, the dc offset can force the output to its limit, and no signal will appear at the output.

The two input terminals of the 741 are connected internally to the two base leads of a bipolar transistor differential amplifier, **fig. 2**. Therefore, some transistor base biasing current must flow through each input terminal of the 741. This input bias current can be as high

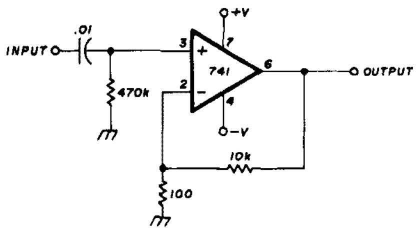

fig. 3. 741 op amp with feedback network selected for a gain of 100. Dc output offset of this circuit will be unacceptably high because the inputs see a large difference in resistance.

as 0.5 μA in the 741, but may be typically 0.1 μA or less. Ideally, the two input bias currents should be equal, but due to differences in transistors, they may differ by as much as 0.02 to 0.2 μA. This difference in input bias current is called input offset current.

To see how this input bias current can effect output offset voltage, consider the circuit of fig. 3. This amplifier is designed to have a voltage gain of 100 and an

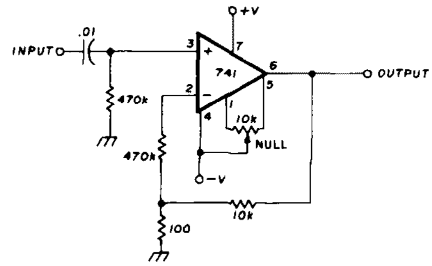

fig. 4. Both input terminals of this 741 circuit see about the same resistance, and output offset may be nulled to zero. Gain of the circuit is 100.

input resistance of 470k. Power supply and null circuits are omitted for clarity. Suppose the input bias current for both input terminals is 0.1 μA. The voltage drop across the 470k resistor connected to pin 3 will be 47 mV. At pin 2, however, most of the input bias current will flow through the 100 ohm resistor, because that is the path of least resistance. The voltage drop caused by 0.1 μA flowing through 100 ohms is only 10 μV, so the input bias current flowing through these resistors produces a dc voltage difference between pins 2 and 3 of nearly 47 mV. Since the gain of the amplifier is 100, the dc output level of the 741 will be 100 times 47 mV, or 4.7 volts under no-signal conditions. This is hardly a desirable situation.

The offset null adjustment pot (shown in fig. **1**) can only compensate for about 15 mV, typically, of input offset, so it would be unable to correct for the 47 mV offset and bring the output level of the 741 back to zero.

To avoid such problems as this, it is good practice to arrange the circuitry so that each input terminal of the 741 sees approximately the same amount of resistance; this will minimize input offset due to input bias current.

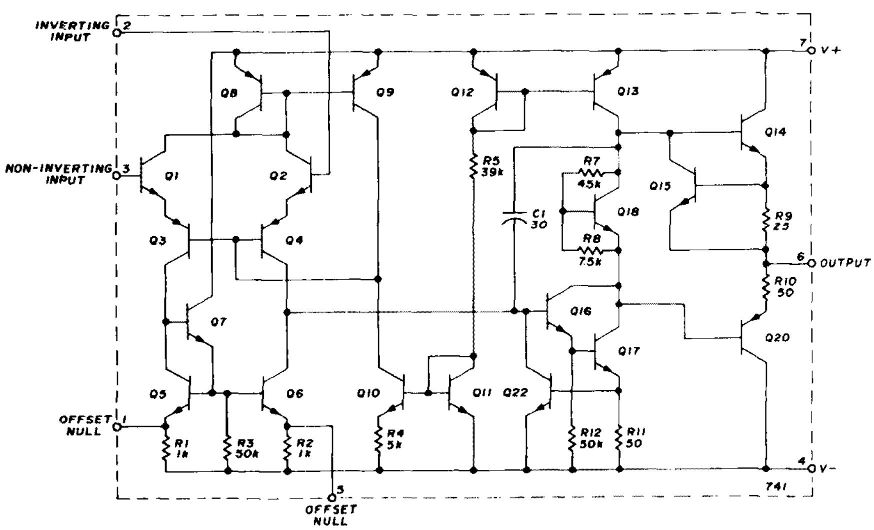

fig. 2. Internal circuitry of the 741 frequency-compensated operational amplifier.

Fig. 4 shows how a resistor may be added to accomplish this. With this circuit, the dc output level may be zeroed with the null pot, and voltage gain is still 100. Keeping all input resistors as small as possible also helps to reduce input offset, but when a high input resistance amplifier is needed, this is not feasible. In calculating the resistance that each input terminal sees, find the equivalent resistance that all the resistors connected to that input terminal would have if they were connected in parallel.

Another approach to the dc offset problem is to reduce the dc gain of the amplifier to unity while maintaining the ac gain at 100. Fig. 5 shows how to do this. Here the dc input offset due to input bias currents of 0.1 μA is 46 mV, but the dc gain of the amplifier is unity, so this will result in only 46 mV offset at the output. Such a small dc output level is insignificant in many cases, and the null pot would be omitted. The 5 μF capacitor should be a non-polarized type. Its value was selected to have a reactance of 100 ohms at 300 Hz, to produce a lower 3 dB cutoff point for the amplifier at this frequency.

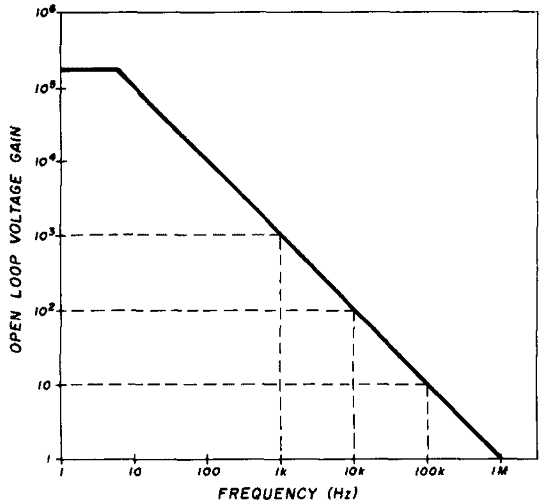

fig. 6. Open-loop frequency response of the 741 operational amplifier.

bandwidth and slew rate

Fig. 6 shows a typical plot of the 741's open-loop voltage gain vs frequency. According to this graph, if

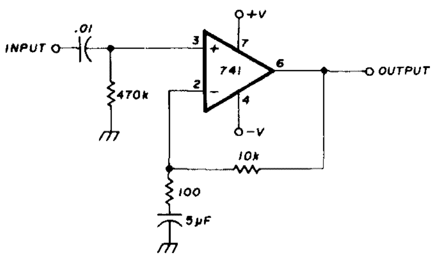

fig. 5. In this 741 circuit the dc gain has been reduced to 1.0, reducing offset problems, but ac gain is still 100.

you designed the feedback network so the op amp had a gain of one, the bandwidth would be 1 MHz. For a gain of 10, the bandwidth would be 100 kHz; with a gain of 100, the bandwidth would be 10 kHz, and so on. That's all well and good, but it's not the whole story when considering gain vs bandwidth tradeoffs.

Slew rate is another parameter of the 741 which requires consideration at the higher frequencies. It is a measure of how fast the output voltage can change; for the 741 it is typically 0.5 volt-per-microsecond. This means that if you design a 741 op amp circuit to have a voltage gain of one, its bandwidth will effectively be 1 MHz only if the signal level is kept small enough to comply with the slew rate limitation. Suppose you feed a 5-volt p-p sine wave signal into this gain-of-one amplifier and start increasing the frequency. At low frequencies, the output voltage would be 5 volts p-p, but by the time you get to 1 MHz, the output would be down to the order of 0.25 volt p-p. This is quite a drop with the input still at 5 volts p-p. Thus the bandwidth is effectively much less than 1 MHz, even though the gain is only one. Moreover, the apparent bandwidth will be a function of how large a signal you use to measure the bandwidth!

Fig. 7 shows a typical curve of 741 output voltage swing as a function of frequency. Below 10 kHz, output swing is determined by power supply voltage; above 10 kHz, however, output voltage swing falls off rapidly due to slew rate limitations.

Of course, slew rate directly controls rise and fall time in non-linear applications. If the power supply is two 9-volt batteries, then the output of a 741 would typically be able to swing between -7 and +7 volts. At a slew rate of 0.5 volt-per-microsecond, it would take 28 microseconds to rise or fall between these two levels.

Simple audio amplification is the most obvious use for 741 op amps, and reference 1 covers most of the other applications which come to mind. Fig. 8 shows how simple gain blocks can be cascaded to provide any desired amount of gain. In order to obtain a 10 kHz bandwidth for each 741 op amp, its gain is set at 100, which is equivalent to 40 dB. When two amplifiers hav-

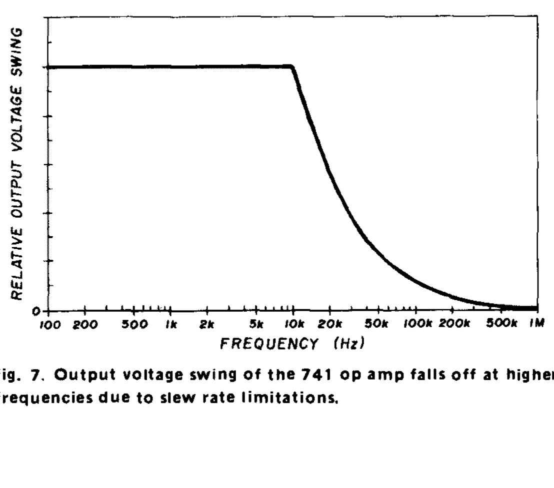

fig. 7. Output voltage swing of the 741 op amp falls off at higher frequencies due to slew rate limitations.

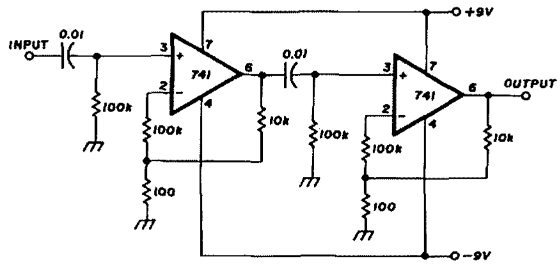

fig. 8. In this circuit two 741 op amps have been cascaded to provide 80 dB of audio gain with bandwidth of about 300 Hz to 6 kHz.

ing 10 kHz bandwidth are cascaded, the overall bandwidth drops to about 6.4 kHz; bandwidth of three stages would be 5.1 kHz. Null pots are not used as the dc output of each amplifier is typically less than 1.0 volt. Capacitor coupling should be used between each stage, however, to prevent the dc offset of the first amplifier from being amplified by the second.

If volume requirements are not too great, 741 op amps will drive a speaker at a comfortable room level. **Fig. 9** shows a simple means of matching the output of the 741 to an 8-ohm speaker. If the peak-to-peak audio swing at the output of the 741 is 12 volts, the power into the primary of the output transformer will be about 36 mW. With larger, more efficient speakers, this is adequate for many applications. The 1.0 μF output coupling capacitor provides low-frequency cutoff of about 300 Hz. This capacitor should be a non-polarized type.

conclusions

One ground rule for using 741 op amps is to make both input terminals see the same amount of dc resistance. Another is that audio bandwidths for voice communication require that each 741 op amp stage have a gain not much more than 40 dB. Also, don't expect much output signal swing above 10 to 30 kHz.

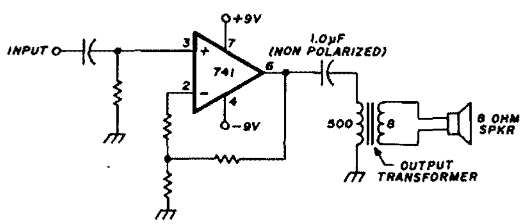

fig. 9. Method of coupling a 741 op amp to a speaker for applications where moderate speaker volume is required.

The 741 op amp is inexpensive, readily available, and easy to use, but, like everything else, it has its limitations. Knowing these limitations will help you decide when and where to use it.

reference

1. Donald Nelson, WB2EGZ, "Operational Amplifiers," *ham radio,* November, 1969, page 6.

ham radio

corner-fed loop antenna for low-frequency dx

Design data for a triangular loop antenna covering 80 through 10 meters

By Barry Kirkwood, ZL1BN, 11 Bromley Place, Auckland 6, New Zealand

For those amateurs who can't erect a beam antenna for one reason or another, a triangular-shaped wire antenna provides a fair compromise. On these pages you'll find details of such an antenna that has given a good account of itself on all the high-frequency bands. It requires little more space than an inverted vee. However, the antenna does require some means of support that is at least 70 feet (21.3m) high. Advantages of the antenna are:

A. Feedline can be coax cable, TV twin lead, or open wire line.

B. Only one support is needed.

C. Vertical radiation angle seems to be quite low, which is needed for DX work.

D. Results on all bands (except 160 meters) appear to be better for long-haul DX than with the sloping dipole, or inverted vee, which is band limited.

triangular antennas

Several triangular wire antennas have been described.[1,2] These are single-band antennas that are variations of the full-wavelength loop. Such antennas, when mounted vertically and excited at the center with second and other even-harmonic energy, radiate straight up -- not the best for DX work.

G3AQC conducted tests of loop antennas close to ground using vhf modeling techniques.[3] He found that full-wave quad and delta loops mounted in the vertical plane with their highest points one-quarter wavelength above the ground, fed symmetrically with the feed point halfway along the base or at the apex, produced high-angle, horizontally polarized radiation and showed little superiority over a simple dipole or inverted vee at one-quarter wavelength height. These configurations are shown in **fig. 1**.

If a delta loop is inverted so that it has a flat top and its apex points down, as in **fig. 2**, a low-angle, vertically polarized lobe appears, which is omnidirectional in the horizontal plane. G3AQC gives details of a practical antenna of this type, which is said to perform well on all bands from 80 to 10 meters and to have a radiation resistance of around 200 ohms.

corner-fed loop

Even more interesting is the result obtained when a delta loop is fed at one end of the horizontal section, as

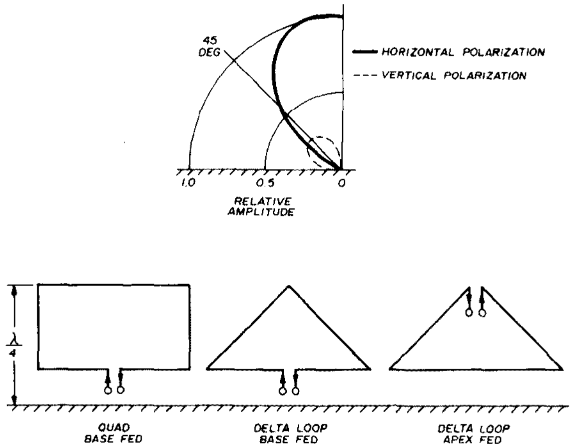

fig. 1. Full-wave loops close to ground. Radiation pattern is little different from a dipole at 1/2 wavelength above ground.

in **fig. 3.** In this case the horizontally polarized radiation is suppressed. The signal is concentrated into low-angle, vertically polarized radiation. The radiation patterns are much the same for both upright and inverted loops; hence the upright delta loop seemed to be the configuration of choice since it required only a single supporting mast.

G3AQC gave no information on the performance of such an antenna on harmonic frequencies. Since the configuration looked attractive, I decided to try such a loop (270 feet, or 82.3m overall) as an all-band antenna.

A triangular 275-foot (83.6m) loop of insulated hookup wire was erected with the apex at 70 feet (21.3m), and the corners of the base were suspended by insulated cords to nearby trees at 15 feet (4.6m). The droop of the base section was within 10 feet (3m) of ground level at the lowest point, which was directly below the apex. Using a simple rf bridge, the length of wire in the loop was adjusted for resonance slightly above 3.5 MHz, where the radiation resistance was 65 ohms. On

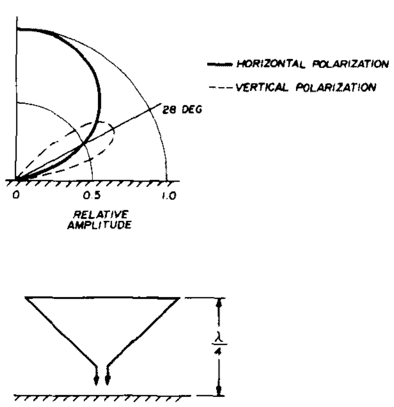

fig. 2. Inverted delta loop at 1/4 wavelength above ground showing vertical radiation pattern in the plane of the loop. Note the low-angle vertical lobe. (After G3AQC.)

the second harmonic (resonance in the 7-MHz band), radiation resistance was about 200 ohms and increased slightly on the higher bands, approaching 300 ohms on 28 MHz.

feed system

The antenna was fed with 300-ohm line. An antenna coupler of the Z-Match type was used to couple the line to the transmitter, as in **fig. 4**. The side of the line connected to the base of the antenna loop was grounded at the antenna tuner, mainly for lightning protection. (Little or no discernible change occurred in loading or performance on transmit or receive with the ground connection on or off.)

G3AQC states that this antenna is fundamentally unbalanced. If it is fed with coax line, the braid should be attached to the horizontal leg of the antenna.

The antenna can be fed with coaxial cable, twin lead, or open-wire line with minimal loss and moderate impe-

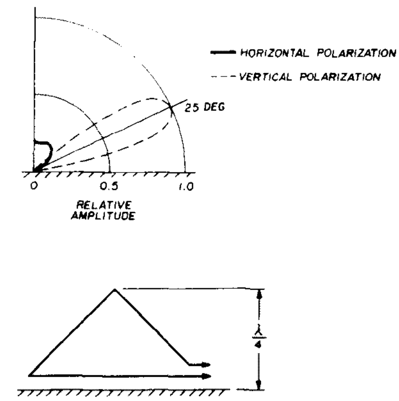

fig. 3. Delta loop, corner fed, showing low-angle, vertically polarized radiation.

dances occurring at the transmitter end of the line. An antenna coupler is recommended, not only for improved efficiency and hamonic reduction but to prevent out-of-band signals from overloading the receiver.

evaluation

Antenna evaluation is difficult because of the variables involved. The task is even more difficult because of the patchy band conditions that have prevailed since the loop was erected. The antenna is located on a typical suburban site, which is cluttered with buildings, small trees, and power lines. A low ridge is between my station and the major DX propagation paths: northwest to Europe and northeast to North America. Comparative results with other amateurs in Auckland when using other antennas at ZL1BN indicate that the location is only fair for DX on these paths, although it is quite good on the long path to Europe. The antenna is oriented along a line 110-290 degrees true, which puts it end-on to Rome and Lima and broadside to Alaska and South Africa.

On 80 meters, DX seems at least as good as with an inverted vee at the same apex height. Too few DX openings have occurred to evaluate directivity, although theoretically directivity should be almost omnidirectional on the long haul. At intermediate ranges, out to about 3000 miles (4800 km) the loop is clearly superior to anything ever used at this station.

On 40 meters performance appears superior to a ground plane used previously. Exceptionally good reports have been received both from Europe and the USA, especially under marginal conditions, which indicates that low angle propagation exists both broadside and end-on on this band.

Reports from South America have been good, but it's possible that radiation is better in the direction away from the point of connection of the feedline. Signal strengths from the loop run about 3 dB below antennas of nearby amateurs using Yagis at 40 to 60 feet (12-18m) above ground. Signals are weaker broadside to the plane of the antenna. The answer might be to suspend a bisquare array or quad loop for 20 meters in the plane of the big loop. This combination could give good coverage on 20 meters and would be inexpensive and easy to install.

Poor band conditions have prevented an adequate evaluation of performance on 21 and 28 MHz. The relatively few contacts made indicate that the pattern is similar to that on 14 MHz with broad lobes off the ends of the loop and nulls on the sides.

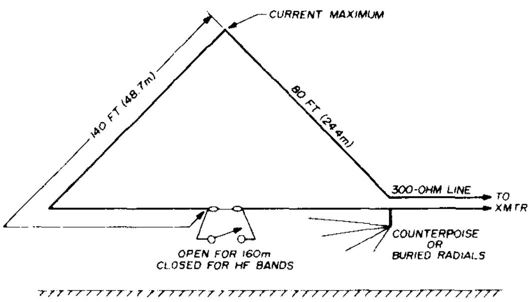

fig. 5. Possible conversion of loop for 160-meter DX operation. Counterpoise or buried radial system should be as extensive as possible. Fanout of ground system below antenna will assist operation on all bands.

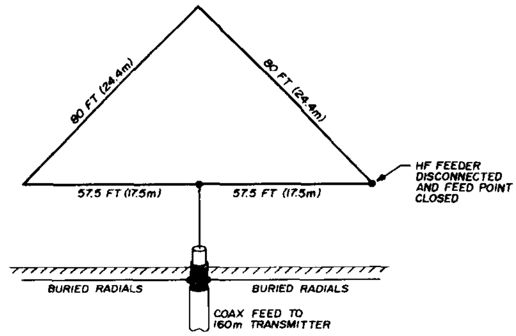

fig. 6. Possible conversion of loop to a vertical monopole radiator for 160 meters.

top band

No attempt has been made to try the loop on 160 meters. If the loop were opened opposite the feed point, it would resonate as a dipole on that band, but results would not be good, since the current point would be close to the ground. It might be better to open the horizontal leg of the loop so that the current point would be at the apex, and the other side of the feeder could be connected to a radial or counterpoise ground, as shown in fig. 5.

Another possibility might be to feed the lower geometric center with a single wire and work it against ground (fig. 6). According to Krause[4] a half-wave loop fed in this way should show true resonance as a vertical quarter-wave antenna against ground without loading coils. All such experiments can be carried out at ground level.

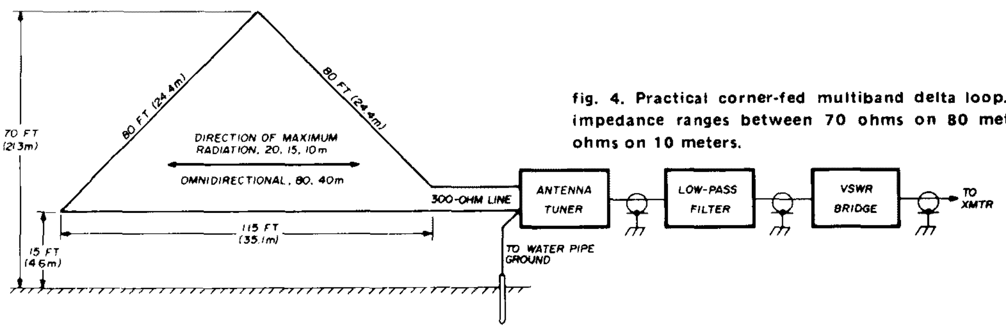

fig. 4. Practical corner-fed multiband delta loop. Feedpoint impedance ranges between 70 ohms on 80 meters to 300 ohms on 10 meters.

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ham radio

1929 - 1941
the golden years
of amateur radio

An informal review
of the technical aspects
of yesterday
that form the basis
of modern
communications technology

Once upon a time, in an era seen through an opaque veil of memory, there existed a Golden Age of amateur radio. Forged in the boom-and-bust year of 1929, the period ran until December 7, 1941, when amateur radio was closed down "for the duration".

These dozen years encapsulated important technical developments that form the basis of amateur radio today. The period opened with the demise of the battery-operated radio and closed with amateur radio into the developmental stages of vhf fm techniques, single sideband, double-conversion receivers and high-gain beam antennas. It was a period of confusion and technical advancement, with the shadows of the great depression and the coming World War always in the background.

It was during the Golden Years that the amateur population of the United States exploded (**fig. 1**). The number of amateurs jumped from 16,829 in 1929 to 54,502 in 1941, the greatest growth being in the years between 1929 and 1934. Before 1929, the American public was bemused by radio broadcasting, which reached the proportions of a craze. New radio stations were coming on the air daily, battery receivers and do-it-yourself kits were popular, and the general public adopted the new sport of radio listening much in the manner of later fads such as miniature golf and the hula hoop. However, by 1929 the broadcast craze was over. The battery radio had given way to the ac operated receiver, which was rapidly becoming a piece of household furniture instead of a seven-day wonder, and commercial broadcasting was a way of life. Until then amateur radio had been overshadowed in the public eye by broadcast listening (an immensely larger hobby) and had remained an esoteric retreat for a few dedicated individuals and eccentrics -- a compact group shielded from the general public by their reticence, and by a widespread interest in broadcast reception.

Gradually the general public became aware of short-wave radio, due partly to the publicity given radio amateurs who pioneered the advance into the shorter wavelengths and led the way on long-distance, high-frequency communications. About the same time, international short-wave broadcasting was started by a few pioneering stations and the American public was thrilled to hear the sounds of Big Ben in London rebroadcast across the country via a transatlantic short-wave relay.

By Bill Orr, W6SAI, EIMAC Division of Varian, San Carlos, California 94070

At the depth of the great depression in 1932, with over one-third of the work force unemployed, young men with idle time on their hands discovered amateur radio and the exciting short waves. Almost overnight short-wave listening and amateur radio caught the public eye and boomed. Short-wave reception, a quick fad, was featured in weekly newspaper columns and the new radios of 1933, in their pristine cathedral-shaped cabinets, boasted at least one short-wave band on their multi-colored dials.

the radio amateur of 1930

The radio ham of 1930, probably unemployed, with little money and plenty of spare time, faced the serious problem of getting on the air at little or no cost to his flat wallet. His technical background was that of a high-school graduate, or a graduate of an electrical correspondence or trade school. Only a small percentage of amateurs were college engineering graduates. Many of the older radio amateurs, of course, worked in the radio industry or in broadcasting, but the just-licensed amateur did not seem to follow the pattern set by his elders.

The task of assembling a ham station was formidable. While many radio stores existed — many more than today — credit was unknown and all purchases were for cash. The price of components and equipment was startlingly high considering the state of the economy. A high school graduate would be pleased to get a job paying $5 a week for 50 hours of work, and radio technicians, on the average, earned from $20 to $30 a week. Most radio components and tubes, however, cost more in 1930 then their modern counterparts do today! It was not until

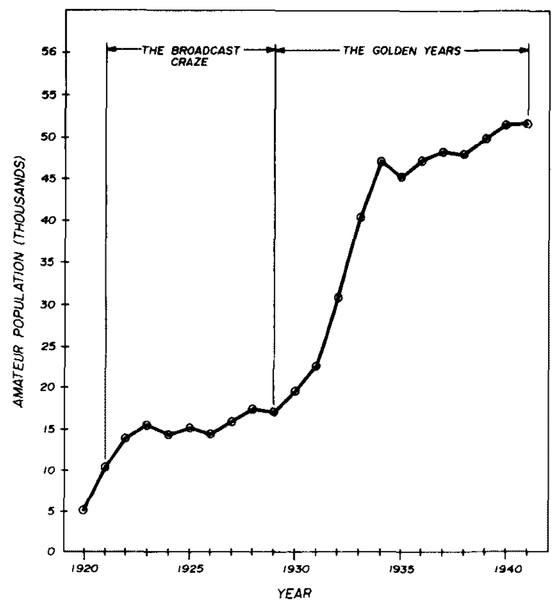
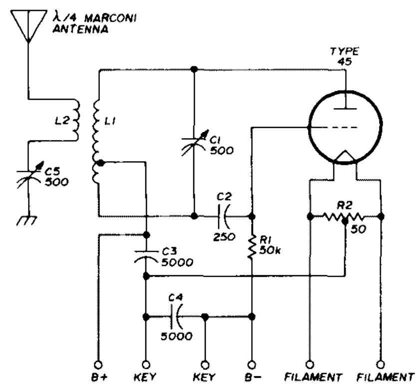


fig. 1. Amateur radio experienced five years of explosive growth between 1929 and 1934, the "population" going from 17000 to 47000. This was due, in part, to the end of the broadcast craze and the introduction of the ac-operated, entertainment broadcast radio. All through the 1920s the American public had a love affair with broadcasting and building broadcast receivers was a popular pastime. When the great depression arrived in 1929 many young men found themselves unemployed with plenty of time on their hands. Short-wave radio provided the next great interest, and amateur radio grew accordingly. By the start of World War II, the U.S. amateur population stood at about 53000.



C1,C5	500 pF variable (Cardwell)
C2	250 pF mica
C3,C4	5000 pF mica
L1	10 turns no. 12 (2mm) enamelled wire on 2-1/8" (5.4cm) diameter bakelite form. Winding length is 2½" (6.4cm)
L2	Same as L1. Adjust spacing between coils for proper coupling
R1	50k, 2-watt IRC metallized resistor
R2	50-ohm wirewound resistor, center tapped

fig. 2. The series-fed Hartley oscillator. Operating in the 80-meter band with a plate supply of 300 volts, this transmitter provides a 7-watt signal. In addition to the components listed above, also needed are five Fahnestock clips, a 4-prong tube socket, and two brown "beehive" insulators to support the coil. Antenna is connected to a series-tuned circuit which is inductively coupled to the oscillator coil.

about 1934, when the radio industry geared up for large scale production of inexpensive household radios, that the cost of components dropped sharply.

Nevertheless, the newcomer was not without sources of radio parts. He scrounged from his friends, haunted the dusty back room of the local radio repair shop for castaway battery receivers, and took an occasional trip to the "radio row" of the nearest big town. With a free, broken-down battery receiver, a few used tubes and a cheap B-eliminator power supply he was ready to build his station and get on the air. Many of today's old timers remember this adventure well. This is the story of how it was done, and the results.

1930 beginner's transmitter

The development of the amateur receiver will be discussed at a later date, but here we'll concentrate on a representative beginner's transmitter of the early 1930s. One or two basic circuits were widely popular at that time, circuits that were easy and inexpensive to build

and sure-fire in operation. Both were single tube, oscillator-style transmitters.

The transmitter was designed around available parts and a good starting point was the readily obtainable, defunct battery receiver of the late twenties. During the lean years of the early 1930s the great majority of amateurs worked CW with a power input of 50 watts or less and many amateurs enjoyed operating with a power level of only 5 or 10 watts. With luck, then, a beginner could join this group and have a lot of fun. The old battery receiver could be torn down for parts, and it was easy to buy a second-hand B-eliminator power supply that, for a dollar or less, would provide about 180 volts at 40 milliamperes. That would suffice for a 5-watt transmitter and a lot of amateur stations all across the U.S. could be worked with that power and a good antenna!

For a few more dollars a 300-volt power supply could be assembled from junk parts to provide the amateur with a *real* transmitter – upwards of 20 watts input. The possibilities were limitless!

One of the most popular transmitter circuits of the time that was passed hand-to-hand among the newer amateurs was the simple series-fed Hartley oscillator, a simple one-tube transmitter that worked well with many of the triode receiving tubes then available for a few cents (**fig. 2**). The whole transmitter used only one tube and eight parts including the tube socket! This rugged and reliable circuit used inductive feedback coupling between the grid and plate of the tube to sustain oscillation. A very high-C tuned tank circuit minimized the effect of capacitance changes and provided dynamic stability to the oscillator. The simple components were firmly screwed to a heavy board which was isolated from the operating table so that any vibration caused by manipulation of the key would not be imparted to the oscillator. When run from a power supply that had reasonably good regulation, and used with a taut antenna that did not swing in the wind, the little transmitter sounded as good (or better) than some of today's more sophisticated equipment.

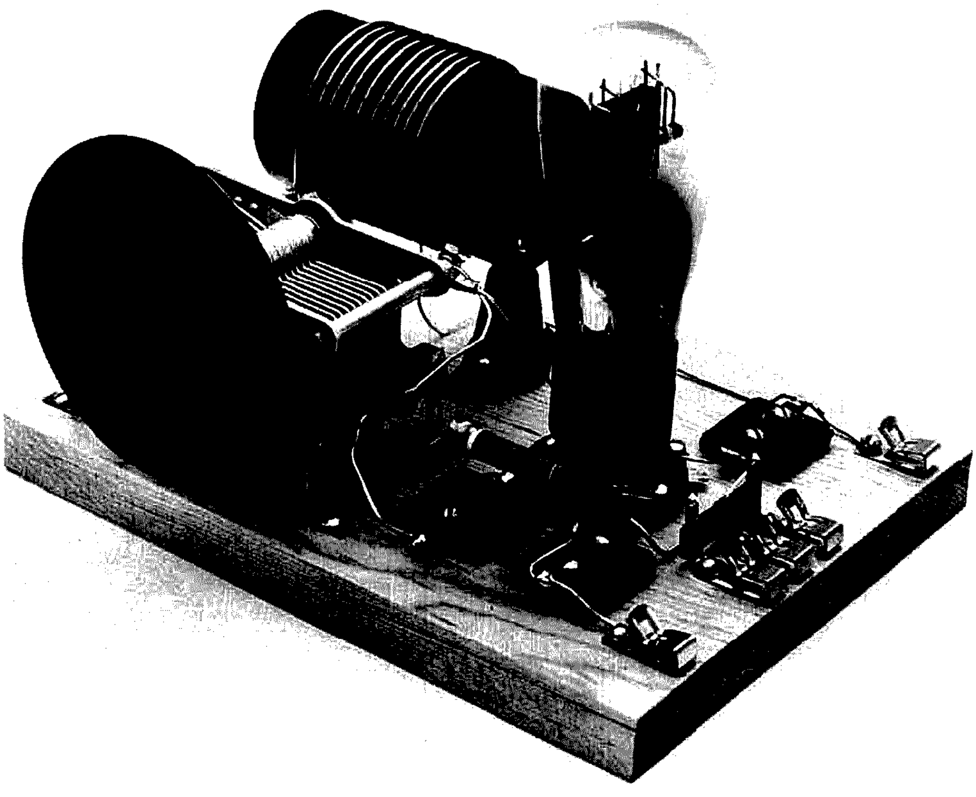


fig. 3. The way it was done in 1930. Using a single 45 tube, this little transmitter gives a good amount of itself in 1976. Built of parts taken from a defunct battery receiver, the Hartley transmitter cost little or nothing to build. High-C circuit provides good stability and when run from a regulated 300-volt power supply, the rig provides a good 1976-style signal. A 160-meter version of this transmitter is shown in the May, 1932 issue of *QST*.

A re-creation of the famous 1930s Hartley transmitter is shown in the photograph of fig. 3. Built in 1976 with hard-to-find 1930 components, this rig has been on the air and has been used to work a number of stations on 80-meter CW. Using a 245 triode at 300 volts and 50 milliamperes plate current (15 watts input), the transmitter puts out a solid 7 watts with good stability. The 245 was a popular tube because it could often be obtained free, for a few cents used, or as a manufacturer's "second" for 39 cents.

A *Cardwell* receiving-type variable capacitor is used, and the tank coil is wound on a genuine bakelite form. The bypass capacitors are uncased mica units, rugged and reliable, but already going out of style as bakelite-cased capacitors came into vogue in 1930. Time to assemble and test the transmitter is about three hours, including giving the "breadboard" a coat of shellac.

The transmitter is keyed in the filament return circuit. Since no waveshaping is included, the keying is *hard* so the resultant waveform may distress a nearby amateur who is operating close to the frequency of the midget transmitter. Any attempt to include a keying filter should be approached with caution as softening of the keying tends to place a "yoop" on the signal. A slight amount of filtering, however, can be used to advantage without disturbing the crystal-like note of the transmitter.

The rf plate impedance of the 245 oscillator is about 3000 ohms. The L/C ratio of the plate tank circuit provides a Q of about 30. The circulating current in the tank circuit, therefore, runs about 6 amperes. The tank coil and leads to the variable capacitor are made with number-12 (2mm) wire to carry this amount of current. The oscillator runs well into the class-C region as the rms grid voltage is close to 140 volts. The μ of the 245 is 3.5 so a cutoff voltage of about 85 volts is required at a plate potential of 300 volts.

The simplest antenna for the little transmitter is a 66-foot (20m) end-fed Marconi, working against ground. A series-tuned circuit, consisting of a coil and capacitor (whose values are approximately the same as those in the main tank circuit) can be used. Alternatively, the same series-tuned circuit can be used to match into a coaxial feed system for a dipole or inverted-V antenna. At W6SAI, a simple 66-foot (20m) wire is used for contacts up and down the West Coast.

transmitter power supply

A bonanza existed for the penny-pinching 1930's amateur in the flood of obsolete B-eliminator power supplies which were rendered useless by the advent of ac operated broadcast receivers. These units provided up to 180 volts dc at 30 to 50 milliamperes and were intended as a replacement for the messy and short-lived B-battery.

Reaching the market in quantity about 1926, the B-eliminator solved the rectifier problem by side-stepping the vacuum diode, going instead to a unique gas rectifier which required no filament. The *B-H tube*, designed and manufactured by the Raytheon Company was the answer, and multitudes of these tubes were available to the 1930 amateur for as little as five cents apiece in the storage bins on radio row (**fig. 4**).

The B-H tube operated on the principle of electron conduction in a gas. Basically, when a potential difference exists between two cool metallic surfaces separated by a gas, the few free electrons in the gas move toward the positive (anode) terminal at a rate which is dependent upon the potential gradient. As the gap is decreased, the gradient is increased to a point at which the electrons attain sufficient velocity to knock an electron off a gas molecule, thus multiplying the number of negative particles. The gas now becomes a conductor.

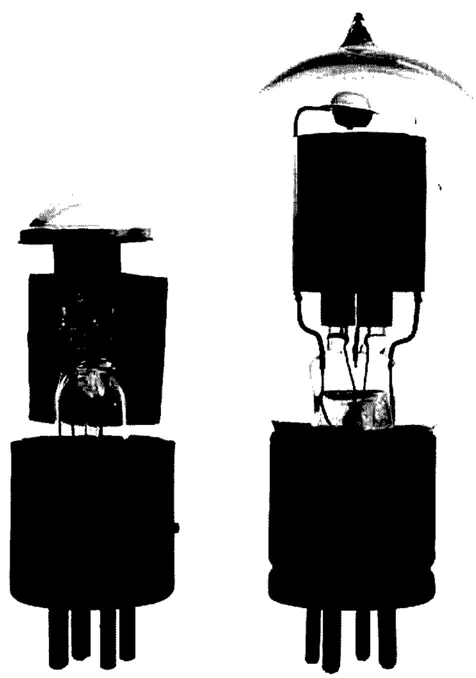


fig. 4. An early full-wave rectifier: the Raytheon BH tube. Designed for B-eliminator service, the BH tube provided up to 180 volts dc at a maximum current of 50 milliamperes. A late production BH tube is shown at left, with an early, brass-based predecessor, the B tube at right (the H standing for "heavy duty"). The larger B tube was rated at 30 milliamperes. Both tubes were gas types with no filament. Shortly after the BH tube was designed it was made obsolete by the 280 (later type 80) vacuum rectifier.

When one electrode is larger than the other, the current will flow from the large electrode to the small one, and when one electrode is point-size, the current flow is practically unidirectional. This means that the anode must be very small in size compared to the cathode and, unfortunately, the anode dissipation is high per unit area. As a result anode material and insulation assume critical proportions.

The problem of building a gas rectifier tube that would work, and deliver reasonable life, was finally solved by a research team under the direction of Dr. Vannevar Bush at the Massachusetts Institute of Technology in 1924.* A special, heat-treated alloy was used for two pin-point anodes, which were surrounded by the larger cathode (**fig. 5**). The "short-path principle" was

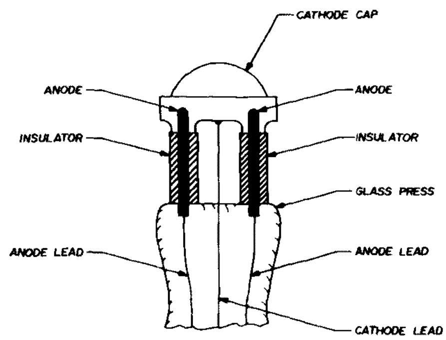


fig. 5. Interior of the BH gas rectifier. In a gaseous atmosphere electrons flow between two conducting points. If one electrode is larger than the other, the current flows from the large electrode (negative) to the small electrode (positive). With a point-type anode the reverse current flow is very small but anode dissipation is severely limited by the small area of dissipation. The "short-path" principle (discussed in text) is used to insure that electrons strike the tips of the anodes. Special, high-temperature lava insulators are used to insulate the tips from the cathode "hat" and glass press.

used to provide a good insulating ring between the anodes and the cathode. This postulate states that a rarefied gas is an insulator between two areas in close proximity if other *points* exist in greater proximity. Thus, to provide a good insulating ring around the anodes and the cathode, the hat-shaped cathode was shaped so as to completely pass around the cathode supports to utilize the "short path" principle as an insulator. Since the discharge cannot pass between the areas in close proximity, only the points of the anodes are struck by the electrons, leaving the insulating material far enough away from the anode heat and the possible disintegrating effect of high temperature. A complete description of the BH tube appeared in the November, 1925, issue of *QST*.[1]

Viewing those days from these days, it is difficult to see why this *round-about* approach was taken to develop a full-wave rectifier as diode conduction had been known for many years and various companies were manufacturing vacuum diodes as rectifiers for telephone equipment.

*The BH tube was developed from pioneering work in the field of gas conduction done in 1916 by Charles G. Smith of Raytheon, aided by J. A. Spencer and M. Andre of *La Radio Technique* of Paris, France. By 1926 more than twenty companies in the United States were licensed by Raytheon to manufacture B-eliminator units using the BH tube.

A logical guess is that a patent problem existed in this area. Looking back, it appears that oxide-coated filaments were generally used by the Western Electric Company and thoriated tungsten filaments were used by RCA and their licensees. Since, by agreement, the Western Electric Company was not in the home-entertainment radio business, the possibility exists that the oxide-coated filament was denied RCA and one of the licensees (Raytheon) turned to the gas tube as the only alternative.

Regardless of the actual reasons, technical or otherwise, the first ac power supplies sold to the general public in large quantities used the Raytheon B-H tube and these, in turn, were obsoleted by the double diode, oxide-coated filament 280 rectifier tube which shortly became available through some mysterious inner workings of the infant entertainment electronics industry.

on the air

And so the amateur newcomer was finally on the air! The receiver, a two-stage job using a 201A detector and 201A audio amplifier, was also built from a defunct battery receiver and could be run from the same B-eliminator as the transmitter. An old automobile storage battery provided the filament power. Sometimes a pair of 199 tubes would be used in place of the 201As — then a no. 6 dry battery would suffice for filament power. With luck the whole station could be assembled for less than five dollars.

The results? Since most amateurs ran low power in those days, the 7-watt transmitter provided a good, workable signal. Looking back through old copies of *QST* reveals that many amateurs, with a transmitter of this type, worked hundreds of stations on 80 meters including contacts with Canada, Mexico and Alaska. Several W6 stations maintained schedules with New Zealand on 80 meters using 10-watt transmitters of this same general type. Some adventurous amateurs put the little Hartley transmitter on 40 and 20 meters and worked real DX, but the problem of drift and stability became onerous at those frequencies.

Building and operating a transmitter of this type is an adventure in itself: the search for authentic parts, the assembly and test, and on-the-air operation. When you tell a station you're working about the transmitter the usual reaction is one of amazement — amazement that such results can be obtained with such simple equipment. When you work an old-timer who remembers the little Hartley the reunion is dramatic and brings back a flood of memories.

So hats off and a silent salute to the beginning amateurs of yesterday, many of whom used the little Hartley, and many of whom are the engineers, innovators and industry leaders of today.

reference

1. *Miles Pennybacker*, "The Raytheon Rectifier," *QST*, November, 1925, page 38.

ham radio

CIRCUITS & TECHNIQUES

Ed Noll
W3FQJ

low-noise two-meter preamp

There has been increased interest in weak-signal DX on 144-MHz recently, and some amateurs who have limited their vhf activity to fm for the past several years are rediscovering ssb activity on the lower sections of the band. Part of this renewed interest in vhf ssb is a result of the commercial multi-mode vhf rigs which are now available, and some is a direct spin-off from OSCAR. In any event, interest in vhf ssb and CW is growing, and in several parts of the country two-meter "activity" nights are being sponsored by local vhf groups to encourage ssb activity. (In the Northeast the two-meter activity night is on Monday evening, beginning about eight o'clock; calling frequencies are 144.110 and 145.010 MHz.)

While many two-meter operators are using modern solid-state converters, a great number have simply dusted off their old vacuum-tube converters and placed them back in service. Hence there is increased interest in low-noise preamplifiers. The two-meter mosfet preamp shown in fig. 1, which was presented at the New England ARRL Convention last fall by K2RTH, is a straight forward, easy-to-build design with a maximum noise figure of about 2 dB (1.7 dB typical). The complete preamplifier is built on a piece of copper-clad circuit board 4½ inches long by 2 inches wide (11.4 x 5.1cm). The transistor is mounted in a shield, also made from 1/16 inch (1.5mm) thick circuit board, which is 1-3/4 inch (4.5cm) square. The preamp is assembled on the main board and all grounds are made by soldering directly to the copper foil (do not use ground lugs). All 470 pF bypass capacitors are also soldered directly to the copper foil with leads no more than 1/16 inch (1.5mm) long.

After the preamp is assembled and all circuit connections have been checked, connect 12 Vdc (maximum) and measure the current drain. It should be 9 to 11 mA. If higher than this, reduce the supply voltage slightly. If your low-voltage dc supply is not variable, you can install a resistor in series with the dc supply line (between C1 and and power supply) and adjust the resistance value for 9 to 11 mA total current drain.

Install the completed preamp in front of your two-meter converter and adjust the two 10 pF trimmers for maximum signal. Gain of the circuit should be about 18 dB and noise figure is 2 dB maximum, so it will really improve the performance of your receiving setup if you're using an old tube-type preamp or converter. You can check the stability of the preamp by touching the transistor with your finger -- nothing should happen. If there's any effect on the received signal when you touch the transistor, check for a bad bypass capacitor (or excessive capacitor lead length).

audio filters

As was pointed out to me recently by WA4KAC, the lowpass audio filter described by OD5CG in the January, 1974, issue of *ham radio*[1] suffers from a loss of attenuation at frequencies beyond the frequency of "infinite" attenuation. This "hump" behavior is characteristic of m-derived filters.

Using the same number of inductors as the original (three), WA4KAC built the lowpass, pi-section filter shown in fig. 2 from the description given by W8YFB in the August, 1972, issue.[2] In WA4KAC's circuit a four-pole, double-throw switch provides the selection of two cutoff frequencies: 650 or 2000 Hz. The filter capacitors were matched to obtain the response shown in fig. 3. The use of two additional 1 μF capacitors for the CW section of the filter(W8YFB used only the two center 2 μF capacitors) results in a sharper cutoff in the CW position and less loss within the passband. With pi-section filters there is no loss of attenuation beyond the frequency of "infinite" attenuation – response continually decreases beyond the cutoff frequency. Cascading filters results in a sharper cutoff response.

OD5CG's audio filter, however, has the advantages of using smaller values of capacitance and provides sharper frequency cutoff. His m-derived filter has an attenuation of 50 dB at 1.5 times the cutoff frequency (-3 dB point on the response curve), whereas the pi-section filter of fig. 2 has 50 dB attenuation at approximately 1.8 times

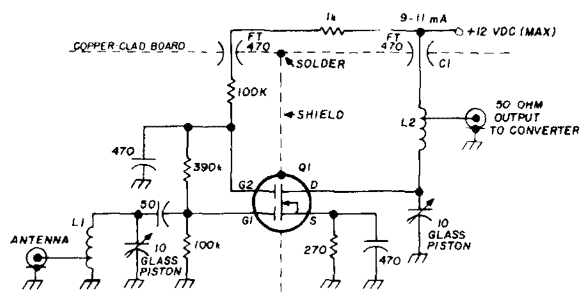


fig. 1. Low-noise two-meter preamp has 18 dB gain and typical noise figure of 1.7 dB. L1 and L2 are each 3½ turns no. 18 (1.0mm) tinned, 3/8" (9.5mm) diameter, 1/2" (6.5mm) long, tapped 1 turn from cold end. Mosfet Q1, in order of preference, is a MEM554C, 3N159, 3N140 or 3N141. These are all unprotected dual-gate mosfets, so use care when handling them.

the cutoff frequency.

A very simple variable audio filter that can be used on both CW and ssb, submitted by KH6AQ, is shown in fig. 4. Although KH6AQ uses this filter with a tube-type audio power stage, a similar arrangement should be suitable for solid-state audio power stages.

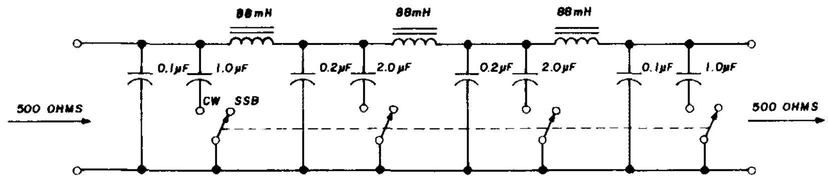


fig. 2. Lowpass audio filter built by WA4KAC has switch for selecting one of two cutoff frequencies: 650 Hz for CW or 2000 Hz for ssb operation. The filter capacitors were matched to obtain the response shown in fig. 3.

dry cell life

Since the power provided by a dry cell is generated by a chemical reaction, and that reaction continues even when the battery is sitting on the shelf, dry cells begin losing power the moment they are produced. Furthermore, batteries discharge faster and faster as time goes on, and the warmer the storage temperature, the shorter the shelf life. Although D-size batteries shouldn't lose more than about 15 per cent of their capacity during the first year after manufacture, smaller batteries (such as the 9-volt transistor radio batteries which are made up of six 1.5-volt cells) may lose 20 to 40 per cent of their charge in the first year. The more expensive alkaline batteries are less sensitive to temperature changes but their shelf life is only a little longer than ordinary carbon cells.

One way to extend the shelf life of dry batteries is to store them in a refrigerator (the chemical reaction slows at lowered temperatures). According to industry experts, new batteries which are placed in a freezer may retain most of their capacity for as long as three or four years. To keep the batteries dry they should be wrapped in plastic before putting them in the freezer — when they are removed from the freezer they should be left in the plastic wrapping until they have warmed up to room temperature. This prevents condensed moisture from forming on the outside of the battery.

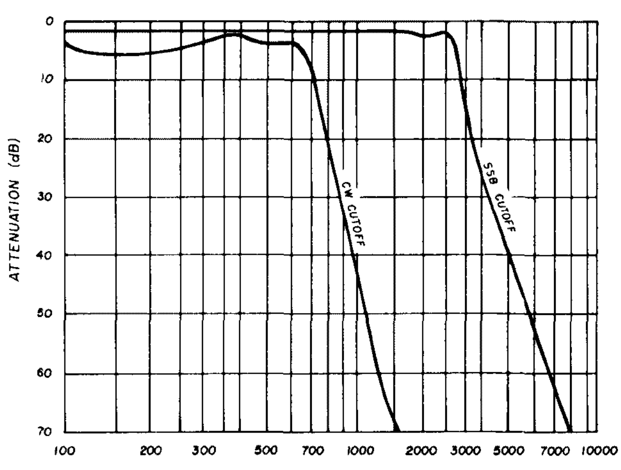


fig. 3. Response curve of WA4KAC's lowpass audio filter shows 650 Hz cutoff (CW) and 2000 Hz cutoff (for ssb operation). Filter capacitors were matched to obtain this response curve.

replacing selenium rectifiers

Although there are direct silicon rectifier replacements for most popular vacuum-tube rectifiers, replacing selenium rectifiers in older electronic equipment poses more of a problem because operating data for these devices is seldom available. However there are two rules of thumb which can be used to determine the correct rating of a silicon diode replacement. The reverse voltage rating of selenium rectifiers is approximately 75 volts per plate. Thus a two-plate selenium rectifier has a voltage rating of 150 volts; three plates, 225 volts; four plates, 300 volts; etc.

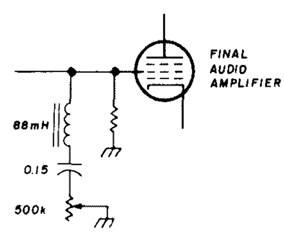


fig. 4. Simple tunable audio filter built by KH6 AQ for use with a tube-type audio power stage.

The current rating of selenium rectifiers is determined by the area of the plates and is given approximately by the relationship

$$current\ rating\ (mA) = 450S^2$$

where S is the length of one side in inches* (assumes square plates). Thus a selenium rectifier with plates which are 3/4 inch (19mm) wide are rated at 250 mA; 1 inch (25mm) wide, 450 mA; 1¼ inch (32mm) wide, 700 mA; etc. This is the maximum current the selenium rectifier can handle at 75 per cent efficiency (the efficiency of seleniums decreases with age). For added reliability, of course, you should choose a silicon replacement with ratings that exceed these calculated voltage and current values.

*In metric dimensions, the current rating in mA is given by $70S^2$ where S is the length of one side in cm.

12 to 6 volt converter

There is still a good deal of used vhf-fm equipment on the market which was originally designed for operation from 6-volt mobile power supplies. Although some amateurs have converted this equipment to 12 volts by rebuilding the power supply, an easier and less expensive solution for medium-powered gear is to build the simple 12 to 6 volt converter shown in fig. 5. With the devices mounted on a suitable heatsink the maximum output current of this circuit is about 15 amperes. If the positive and common lines are isolated from the chassis, the circuit may be used with either negative- or positive-grounded mobile systems.

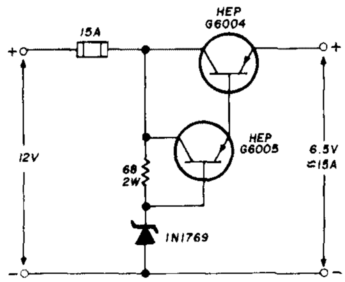


fig. 5. 12- to 6-volt converter suitable for operating low- and medium-powered 6-volt vhf-fm equipment from a 12-volt power supply.

saving your mobile rig

There's little you can do to prevent a thief from breaking into your car and ripping off your mobile rig, but you can slow him down a bit. Burglar alarms, for example, don't prevent thefts – they're only guaranteed to let you know that a theft has already taken place! Since most successful thieves can disarm an alarm in a matter of seconds, commercial auto burglar alarms are little more than an annoyance -- they're certainly not a deterrent. It may be more difficult for a professional thief to disarm a homebrew alarm that he knows nothing about, but since he can break into your car, rip out your gear and be on his way within 20 or 30 seconds, alarms offer little protection. An alarm may scare off an inexperienced thief, but all too often the owner neglects (or forgets) to turn on the alarm when he leaves the car, or doesn't even roll up all the windows and lock the doors.

When you get right down to it, aside from locking your mobile rig in a bank vault, there's very little you can do to completely protect it from departing to points unknown. However, here is a circuit from *The Atlanta Ham* (published by the Atlanta, Georgia, Radio Club) which might help you to recover your mobile gear if it does get ripped off (see **fig. 6**). In almost all cases when the thief leaves your car with your rig he has your microphone, your Touch-Tone pad and the power cord from the fuseholder down. Unless, of course, you put your mike and Touch-Tone pad in the trunk (highly recommended). Usually the rig is simply ripped out, wires dangling. If you build this latch circuit into your rig, however, the thief has a radio that won't push-to-talk when he does and, furthermore, transmits all the time. If you were thoughtful enough to put your mike in the trunk, whenever he attaches power the rig will put an unmodulated carrier on the air, making it easier to track him down. This is especially true in metropolitan areas where local fm operators have been advised to be on the lookout for your "speechless" fm transceiver.

If the circuit is carefully installed in your rig, and the new wiring is worked into the existing wiring harness, it's unlikely that the thief will be able to locate the trouble. And even if you don't recover the rig you may get some satisfaction out of the fact that the thief couldn't get any use out of it and had to junk it!

While we're on the subject of mobile radio gear, are you absolutely sure that your rig is covered by your auto insurance? Some amateurs who have lost their equipment to thieves discovered after the fact that the loss was not completely covered. With many mobile rigs now costing upwards of $500 or more, it's comforting to know that any loss is completely covered. If you're not sure, check with your agent, and make sure the extent of the coverage is in writing. You may be surprised to learn that your policy has only limited coverage – expensive items such as mobile radios and cameras may require additional coverage to be fully insured against loss.

european semiconductors

I have had a number of requests for information on the system used for numbering European semiconductors. Unlike the 2N-system used in the United States, the European numbering system gives a good deal of basic data without resorting to a transistor data book. In their system the first letter indicates germanium (A) or silicon (B), the second letter gives the general construction or application, and the remainder is the device serial number. For the second letter, C indicates an audio type (not power); D, audio power; E, tunnel diodes; F, small-signal rf; L, rf power; P, photosensitive; R, controlling and switching (not power); S, switching transistors (not power); T, switching and controlling devices with specified breakdown characteristics; U, power switching transistors; Y, power diodes; and Z,

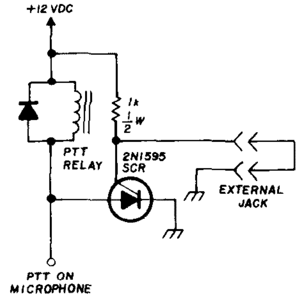


fig. 6. SCR latch turns on your mobile rig when power is applied if external circuit is broken when the rig is stolen. External wires can be run under the dash so thief must cut them when removing the rig from your car. If the PTT relay doesn't have a protective diode across the coil, install one. The SCR is a 100 PIV, 1 amp device such as the 2N1595. SCRs such as the HEP R1003 (0.8 amp) or R1217 (4 amps) should also be satisfactory.

zener diodes. For the serial number, three numerals are used for entertainment types; one letter – X, Y or Z – plus two digits indicate industrial types. The AF117, for example, is a small-signal germanium rf transistor for entertainment purposes; the BCZ11 is an industrial silicon audio transistor.

original which used the optical coupler. In his circuit W2CQH also used a short piece of miniature coaxial cable for the 1-turn output link – the outer shield is grounded only at the coaxial connector so the braid acts as a Faraday shield, eliminating any capacitive signal (and noise) pickup from the circuit.

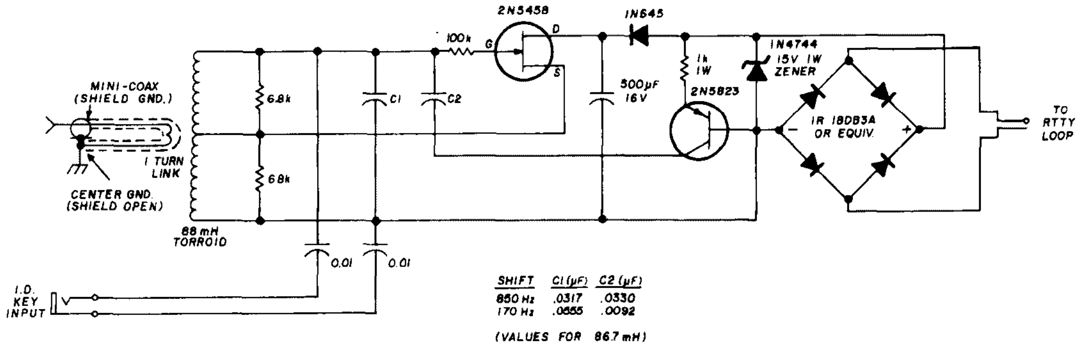


fig. 7. W2CQH's modification of W2LTJ's simple audio-frequency keyer uses silicon pnp transistor switch instead of an optical coupler. Shield on 1-turn coaxial loop acts as Faraday shield.

audio-frequency keyer for RTTY

Several readers have had difficulty obtaining the MOC1002 optical coupler which was used in the simple audio-frequency RTTY keyer described in the August, 1975, issue of *ham radio*.[3] W2CQH faced this problem and solved it with the silicon pnp transistor switch shown in **fig. 7**. Author W2LTJ has also built this version of the circuit and reports that it works as well as the

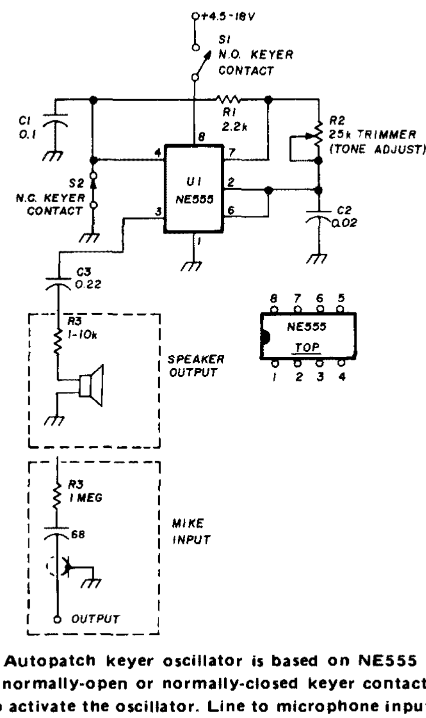


fig. 8. Autopatch keyer oscillator is based on NE555 IC timer. Either normally-open or normally-closed keyer contacts may be used to activate the oscillator. Line to microphone input must be shielded.

keyer oscillator

Shown in **fig. 8** is a circuit for a simple keyer oscillator submitted by KH6IHT and KH6IEL which they designed for autopatch use. Since the autopatch system in their repeater has a decoder bandpass from 2980 to 3080 Hz, R2 is adjusted to 3042 Hz for best results. However, R2 can adjust the output tone over a rather wide audio range for other applications. Two output options are available: speaker or microphone (the microphone input line must be shielded). R3 is adjusted for the required output/input level and may be replaced by a variable resistor, if desired. Normally-closed keyer contacts can also be connected between pin 7 of the NE555 and ground. A 9-volt transistor radio battery (NEDA 918) is recommended for the oscillator.

75A4 noise limiter noise

Some time ago W4ZKI mentioned to the editor that the 6AL5 noise limiter in Collins 75A4 receivers tends to be regenerative, contributing unwanted noise to the receiver. Since the noise limiter is not too effective on ssb or CW, and few operators use it, W4ZKI recommended that the circuit be disabled and the tube removed. This information was passed along to W9KNI who solved the problem very neatly by removing the 6AL5 and plugging in a jumper between pins 2 and 7. On my 75A4 this simple modification reduced the no-signal noise level by about 3 dB, a worthwhile improvement.

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ham radio

MICROPROCESSORS

generating input/output device-select pulses

The preceding column in this series on microcomputer interfacing discussed the 16-bit *out* instruction contained within the 78 instruction set of the 8080 microprocessor chip. The *out* instruction comprises two successive 8-bit bytes and can be written in binary notation as*

$$11010011_2 \ XXXXXXXX_2$$

in 8-bit octal code, 323_8 YYY_8

or in 8-bit hexadecimal code, $D3_{16}$ ZZ_{16}

A discussion of how you convert the 8-bit binary code into either octal or hexadecimal code can be found in reference 1. In the above notations, $XXXXXXXX_2$ represents an 8-bit byte that can range in value from 00000000_2 to 11111111_2; YYY_8 represents a three-digit octal code that can range from 000_8 to 377_8; and ZZ_{16} represents a two-digit hexadecimal code that can range from 00_{16} to FF_{16}. A quick calculation will demonstrate that 11111111_2, 377_8, and FF_{16} all represent the same 8-bit binary word.

The choice of a coding system is up to you. The binary code is awkward to write and difficult to remember. Octal code is used in the popular Digital Equipment Corporation PDP-8 and PDP-11 minicomputer software and is easy to remember. Hexadecimal code is a more natural code for an 8-bit binary word and is currently quite popular among microprocessor manufacturers.

We should emphasize the fact that the manner in which you write the code on paper will not affect the way in which the microcomputer will execute a program. Both octal and hexadecimal code must eventually be converted back to binary code, which is stored in successive 8-bit memory locations. The code conversion can be accomplished in several ways, e.g., by hand or by a computer program.

The second 8-bit byte, $XXXXXXXX_2$, in the 16-bit *out* instruction is the *device code* for the output device. As indicated in previous columns, 256_{10} different devices can be addressed with the aid of such a code. The manner in which this is done is shown in full detail in fig. 1, which provides a device decoding circuit consisting of seventeen SN74154 TTL 4-line to 16-line decoder/demultiplexer IC's. Since this is a rather complicated circuit, we would first like to discuss the simpler decoding circuit shown in **fig. 2**.

The SN74154 IC is a 4-line to 16-line decoder that allows you to input any 4-bit binary word ranging from 0000_2 to 1111_2 and select any single output among sixteen different output channels labeled 0 to 15_{10}. G1 and G2 are the *strobe* or *gating* inputs to this chip; when they are both at logic 0, the SN74154 chip is said to be *enabled* – it is operative, and one of the sixteen output channels, that which corresponds to the binary input at pins 20 to 23, is at logic 0. When either G1 or G2 is at logic 1, the SN74154 chip is said to be *disabled* – it is inoperative, and all sixteen output channels are at logic 1 irrespective of the binary input at pins 20 to 23.

The basic trick that the 8080 microcomputer employs is to *enable* the SN74154 chip for a very short period of time, 500 nanoseconds to be exact. This is done with the aid of a negative clock pulse at G1. This negative clock pulse, called $\overline{IN}$ or $\overline{OUT}$ in reference 1 or $\overline{I/O\ R}$ or $\overline{I/O\ W}$ in the Intel Corporation literature.[2] is generated by the microprocessor chip with the aid of some additional circuitry. $\overline{IN}$ and $\overline{I/O\ R}$ refer to the 16-bit *in* instruction, whereas $\overline{OUT}$ and $\overline{I/O\ W}$ refer to the 16-bit *out* instruction that we are discussing here. During this 500 ns period of time the device code appears on the *memory address bus* and can be used as inputs to the SN74154 chip to select a desired output channel.

The memory address bus is a group of 16 output pins on the 40-pin 8080 IC (**fig. 3**). A *bus* can be defined as follows:[1]

Bus. A path over which digital information is transferred, from any of several sources, to any of several destinations. Only one transfer of information can take place at any one time. While such transfer is taking place, all other sources that are tied to the bus must be disabled.

The important point here is that two types of informa-

*The subscript 2 indicates binary notation, a subscript 8 denotes octal code, a subscript 10 designates decimal notation, and a subscript 16 indicates hexadecimal code.

By Jonathan Titus, Peter R. Rony, and David G. Larsen, WB4HYJ

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon, Inc., Blacksburg, Virginia.

tion can appear on the 16-bit memory address bus: a) the 16-bit memory address for a memory location addressed by the 8080 microprocessor chip, or b) the 8-bit device code present in the second 8-bit byte of an *in* or *out* microprocessor instruction, but not both at the same time. The *in* or *out* microprocessor instruction requires 5 μs for execution, and the device code appears only during the last 1.5 μs of this time.

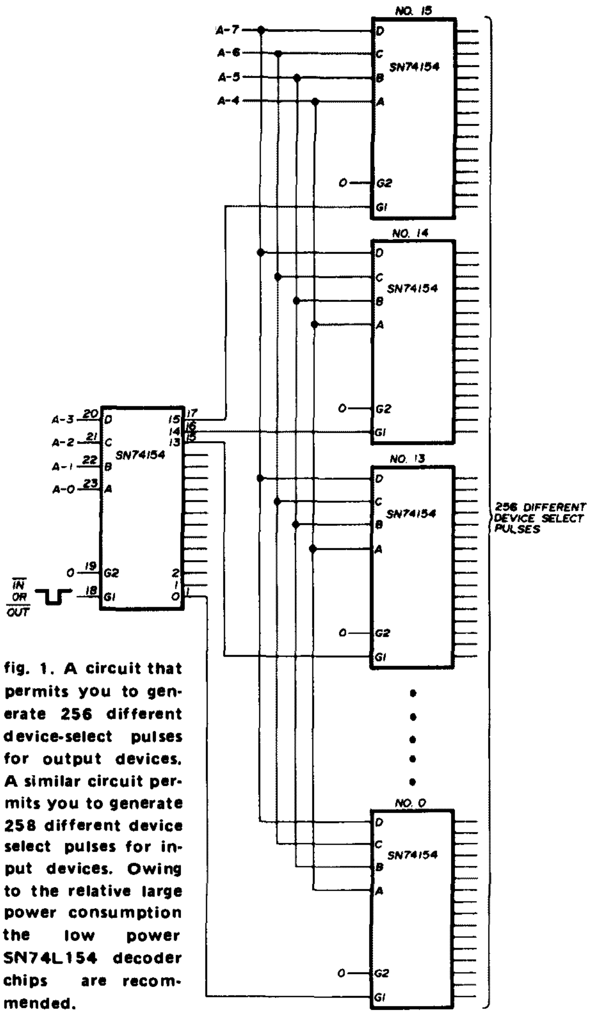

fig. 1. A circuit that permits you to generate 256 different device-select pulses for output devices. A similar circuit permits you to generate 258 different device select pulses for input devices. Owing to the relative large power consumption the low power SN74L154 decoder chips are recommended.

When the device code appears on the memory address bus, the bus is subdivided into two 8-bit bytes, each byte containing the address code. Thus, you have your choice of bits A-0 through A-7, or A-8 through A-15, for the device code. This 8-bit device code is connected directly to one or a group of SN74L154 chips, as is shown in figs. 1 and 2. In fig. 2, only four of the eight device code bits are used, whereas in fig. 1, all eight device code bits are decoded into 256 different output or input device code negative clock pulses.

Each output device is addressed uniquely by the $\overline{\text{OUT}}$ function pulse and a corresponding 8-bit device code. The same is true for each input device; only the $\overline{\text{IN}}$ function pulse is employed instead of the $\overline{\text{OUT}}$ function

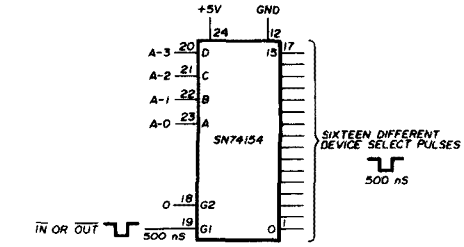

fig. 2. A simpler circuit that allows you to generate sixteen different device-select pulses for either input or output devices but not both simultaneously.

pulse at the gating input G1 to the SN74154 chip. Each device-select pulse lasts for only 500 ns, the time that the SN74154 chip is gated at G1.

Fig. 4 provides a set of timing diagrams that summarizes the external consequences of the 16-bit *out* instruction:

- An 8-bit device code appears on the memory address bus, in this case the code for device 11010001_2 or 321_8, for a period of 1.5 μs.
- During this 1.5 μs, an out function pulse is generated for a period of 500 ns.
- These nine output lines are used as inputs to the seventeen SN74154 IC circuit shown in fig. 1. This circuit generates a 500 ns negative device-select pulse for device 321_8. All the remaining 255 outputs from the decoders remain at logic 1.

This device select pulse can be used to turn on the solid-state relay shown in the circuit in last month's column.

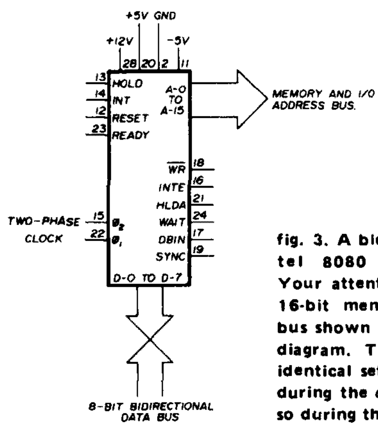

fig. 3. A block diagram of the Intel 8080 microprocessor chip. Your attention is directed to the 16-bit memory and I/O address bus shown at the top right of the diagram. This bus provides two identical sets of 8-bit device codes during the *out* instruction and also during the *in* instruction.

The program, which is analogous to one given previously, is simply:

memory address	octal instruction	description
0	323	Send device select pulse to device given by the following 8-bit device code
1	321	Device code for clear input to SN7474 flip-flop
2	166	Halt the microcomputer

For further information, please refer to last month's column.[3]

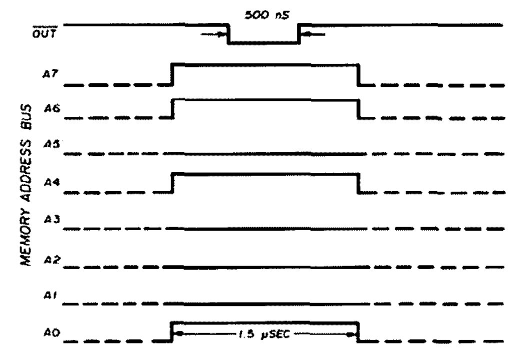

fig. 4. A set of timing diagrams that depict the relationship of the $\overline{OUT}$ function pulse and the device code information appearing on the memory address bus. This information is applied to SN74154 decoder circuits such as those shown in figs. 1 and 2.

In the above paragraphs, we have discussed the interfacing technique called *accumulator I/O*, which is also known as *isolated I/O* in the Intel Corporation literature.[2] A much more exciting interfacing technique is *memory I/O*, which is also known as *memory mapped I/O*,[2] in which an I/O device appears to the microcomputer CPU as a simple memory location. Without question memory I/O will be the most popular interfacing technique among all of the different microprocessor families. One important advantage of this technique is the considerable number of integrated circuit chips that have already been designed for memory I/O applications. Included among such chips are the 8255 programmable peripheral interface, the 8251 universal synchronous/asychronous receiver/transmitter (USART), the MC6820 peripheral interface adapter, and the XC6850 asychronous communications interface adapter. We shall discuss this alternative I/O technique next month.

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ham radio

integrated circuit ssb transceiver for 80 meters

The versatile LM373 IC is the heart of this single-band sideband transceiver that also features some novel diode switching of the signal path

The increasing availablilty and sophistication of integrated circuits has prompted the publication of a number of solid-state transceiver designs.[1,2,3] The transceiver design presented here uses readily available ICs and includes varactor tuning of the variable oscillator and transmitter buffer stages. The receiver's performance is enhanced by a dual fet cascade rf amplifier, and the transmitter provides about 8 watts output from an inexpensive rf power transistor.

circuit

Except for the varactor tuned oscillator (vto), bfo, rf amplifier, transmitter driver and final amplifier sections, all functions are performed by integrated circuits. With the exception of the B+ switching, which is accomplished with a single miniature relay, all other switching between receiver and transmit is done by high-speed 1N914 switching diodes. The transceiver's i-f frequency is 455 kHz, and the vto tunes from 3.045 to 3.545 MHz. Sideband selectivity is provided by a 455-kHz Collins mechanical filter.

Two MPF102 fets in cascade configuration with high-Q toroidal inductors serve at the receiver's rf amplifier (fig. 2). The first design attempts excluded an rf amplifier, but the LM373 used in the front end displayed such poor cross-modulation characteristics and overloaded so easily that an additional high-Q tunable rf stage was necessary. However, even with an rf preselector an occasional strong local station overloaded U1. To protect against this kind of interference a manual second rf gain control, consisting of R1B in a voltage dividing network to provide adjustable bias to pin 1 of U1, was included.

The received signal is fed through relay K1 to the dual fet rf amplifier, through another tuned circuit, and then to U1, a LM373 serving as both the second rf amplifier and the receiver mixer. The vto input is injected at pin 6 of U1 during receive, and the mixer output is then diode switched into pin 2 of U2, the common i-f amplifier (fig. 3).

In addition to serving as the receiver/transmitter i-f strip, U2 also functions as the receiver product detector. Sideband selectivity is provided by FL1, a 455 kHz Collins mechanical filter, inserted between pins 2 and 9 of U2. The LM373 adequately compensates for the filter insertion loss, and it does triple duty as the product detector when the 456.35 kHz bfo output is diode switched onto pin 6 during receive. The audio output of U2 is switched into the audio amplifier, U3, an MFC9010 IC capable of producing two watts of audio output with low distortion.

beat-frequency oscillator

The bfo is a rather conventional bipolar crystal oscillator coupled to a fet source follower by a miniature 455 kHz i-f transformer, T2 (fig. 4). The rf output is made adjustable (through R7A) so that the bfo injection voltage can be set for maximum carrier suppression. The bfo frequency is 456.35 kHz for lower sideband operation or 453.75 kHz for upper sideband.

variable oscillator

The variable oscillator is varactor tuned using a 1N594 diode and incorporates an MPF102 fet source follower buffer (fig. 5). With the values given, the oscillator tunes from 3.045 to 3.545 MHz with the full excursion of potentiometer R8. The +12 volt source for the vto should be well regulated, preferably with a three-terminal 12 volt IC regulator such as the Fairchild 7812. In the transceiver I built, the vto output shifted frequency by a few hundred hertz when switched between receive and transmit, probably due to the difference in load resistances placed on the vto in the two modes. However, by adding R9 and associated circuitry, the receive and transmit frequencies may be synchronized.

By W. J. Weiser, M.D., VE3GSD, 98 Banstock Drive
Willowdale, Ontario, Canada

transmitter circuits

A 741 operational amplifier is used at U4, the speech amplifier (**fig. 6**). A Motorola MC1496 is used as the balanced modulator after attempts with a LM373 proved unsuccessful. The circuit shown does an excellent job, despite its simplicity, and was borrowed from previous designs.[3,4]

The 456.35 kHz carrier oscillator signal is diode switched during transmit to U5, the balanced modulator, and the resulting double sideband output is switched into the common i-f amplifier U2. U2 passes the dsb signal through FL1 and mixes the filtered ssb output with the injected vto frequency. The resultant 3.5 to 4.0 MHz ssb signal is finally switched during transmit to the chain of transmitter buffer and amplifier stages.

to have one available. A rather elaborate zener-regulated bias network was needed to operate this device in a reasonably linear fashion. The collector tuned circuit is similar to that of the driver stage.

The rf output passes through another low impedance lowpass filter which attenuates the harmonic output and also provides an impedance match for a 50-ohm load.

control circuits

All B+ voltages in the transceiver are switched through a single 4pdt relay. One pole switches +12 volts between the receive and transmit switching diodes, and also between the rf amplifier and U1 (on receive) and the speech amplifier, balanced modulator, and transmitter driver amplifiers on transmit. A second set of contacts switches +12 volts from a separate supply to the transmitter final amplifier stage on transmit. The remaining two poles function as an antenna T-R switch and as the receiver offset tuning disable.

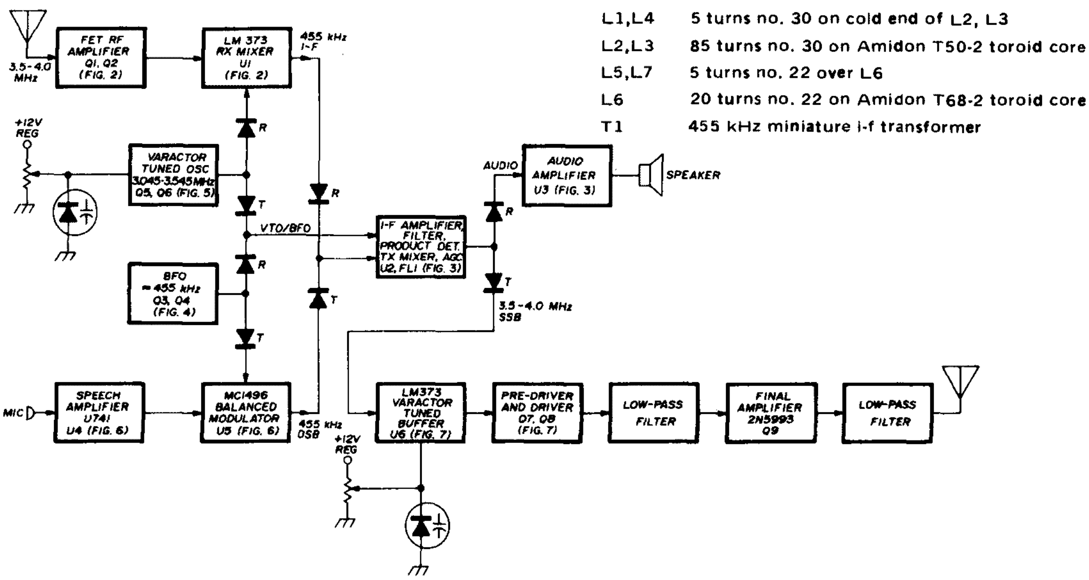


fig. 1. A simplified block diagram of the transceiver. The diodes marked R are biased into conduction during receive; those marked T conduct during transmit. All T-R switching is handled in this manner except the final rf amplifier output. Varactors are used in the tuned circuits of the vfo and buffer stages.

The ssb signal at the output of U2 is extremely weak and requires a number of stages of linear amplification to reach an rf voltage capable of driving the final amplifier, Q9. The output from U2 is first amplified by varactor-tuned buffer U6, also an LM373; the output of U6 in turn excites pre-driver Q7; its output is tuned to the center of the 80-meter phone band by C5 and link coupled through L3 to the driver transistor, Q8, a 2N3553 bipolar transistor (**fig. 7**).

Several transistor types seem to work well in the driver stage (of which the 2N2102 is the least expensive). An adjustable voltage-dividing network utilizing a 3.6 volt zener regulator biases the final into the linear range (**fig. 8**). The driver's output is link coupled via L15 to a 4 MHz lowpass filter.

The final amplifier, Q9, is a 2N5993 rf power transistor. This device is intended for CW and fm use and was reluctantly forced into linear service because I happened

construction

The heart of the transceiver is built on an 8x10 inch (20x25cm) single-sided copper-clad circuit board. The board is pre-drilled and hand etched to incorporate all circuits and components except the vto, audio amplifier, speech amplifier and final amplifier; these are built as small sub-modules on boards which are mounted with L-brackets onto the main circuit board.

The vto is mounted in a small aluminum enclosure. The 2N5993 final amplifier is built on a small circuit board with the rf power transistor mounted on a 2x5 inch (6.3x15.2cm) finned heatsink which is bolted flush with the phenolic side of the circuit board. A shield

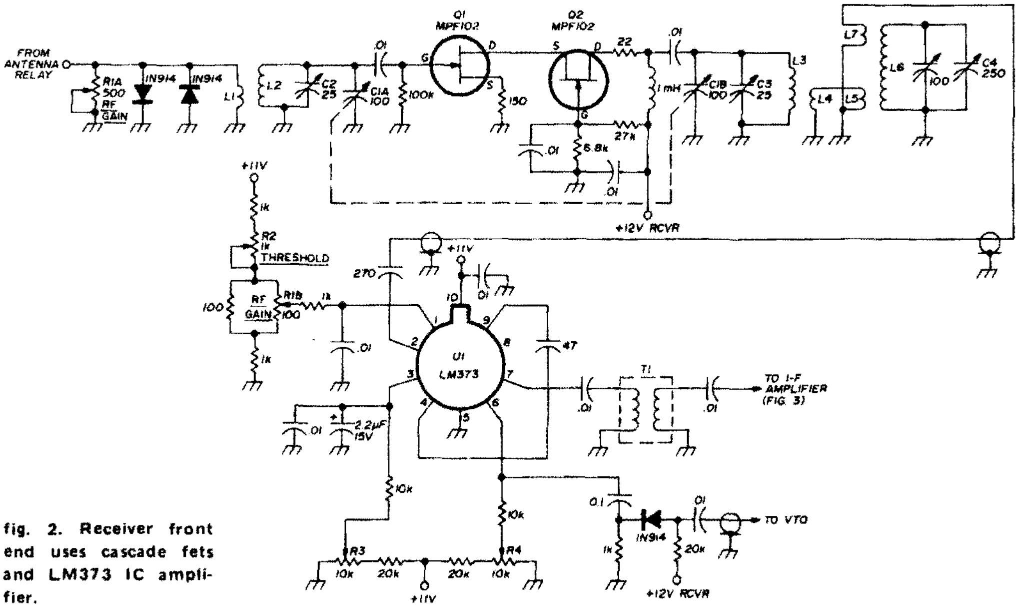


fig. 2. Receiver front end uses cascade fets and LM373 IC amplifier.

separates the final amplifier stage from the main circuit board.

All tuning controls are mounted on aluminum brackets secured directly to the main circuit board's front edge. Finally, the main board is mounted with standoffs to a 10x12x2 inch (25.4x30.5x5.1cm) aluminum chassis. All tuning controls are brought out through the front panel by shaft couplers and universal joints. The output meter is mounted on the front panel; the gear drive assembly was harvested from an old ARC-5 receiver and is mounted on the right side of the panel as the vto tune control.

receiver alignment

The vto was aligned first. Inductor L8 is adjusted so that the desired frequency range of 3.045 to 3.545 MHz occurs at the extremes of rotation of R8. A grid-dip meter and general-coverage receiver were helpful in this

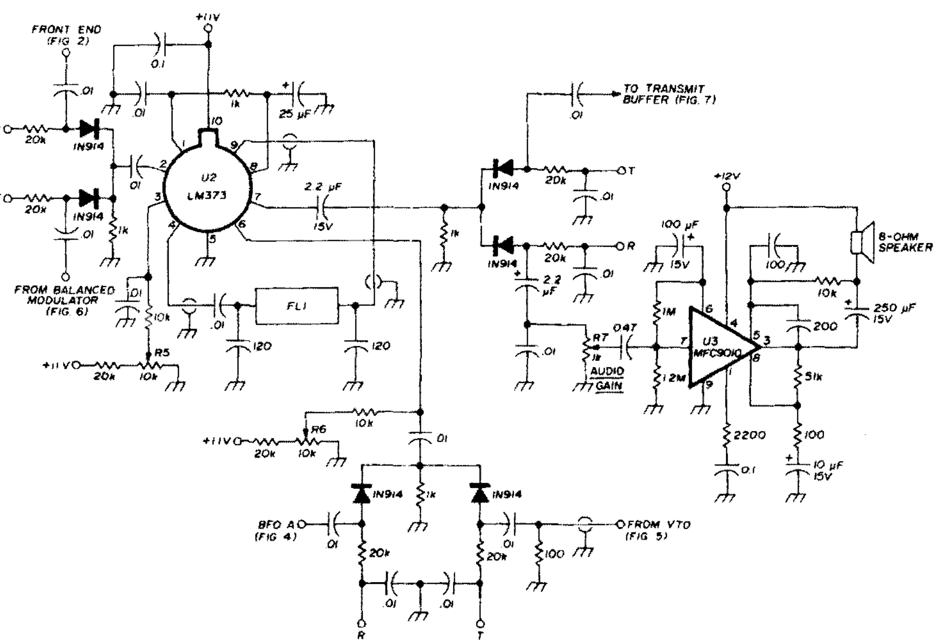


fig. 3. Common i-f amplifier, filter, product detector/transmitter mixer stage. By applying +12 volts to the T or R terminal, the various inputs and outputs are switched by 1N914 diodes. FL1 is a 455 kHz Collins mechanical filter.

alignment. The bfo requires no adjustment and can be easily checked for oscillation with a vtvm and rf probe.

The receiver is easily adjusted by tuning in a strong local signal and alternately adjusting trim pot pairs R3, R4 and R5, R6 for maximum signal strength. Final peaking is accomplished with C1, C4, and the rf tune and receiver tune capacitors.

transmitter alignment

To begin the transmitter alignment, first check the operation of the speech amplifier by monitoring its output with a pair of headphones. Next, loosely couple pin 7 of U2 to a receiver tuned to the transceiver's frequency. A clear, crisp ssb signal should be heard. To null the carrier, adjust R11 for minimum carrier, while at the same time setting R7, the bfo output level adjust, to obtain maximum carrier suppression possible with adequate signal gain.

Shift the receiver coupling to L13, and place a number 42 pilot lamp across that inductor as a dummy load. Set the vto to 3.8 MHz and peak the buffer tune control (R12) and the pre-driver tune control (C5) for maximum signal at the receiver.

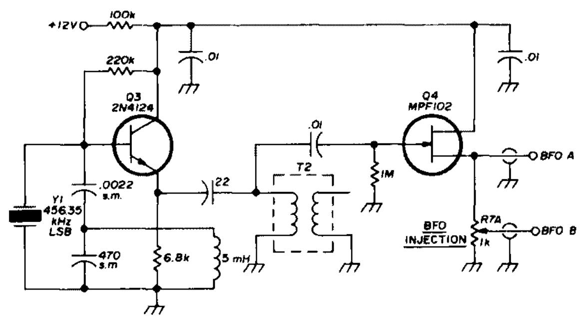


fig. 4. Bfo and carrier oscillator. T2 is a miniature 455 kHz i-f input transformer.

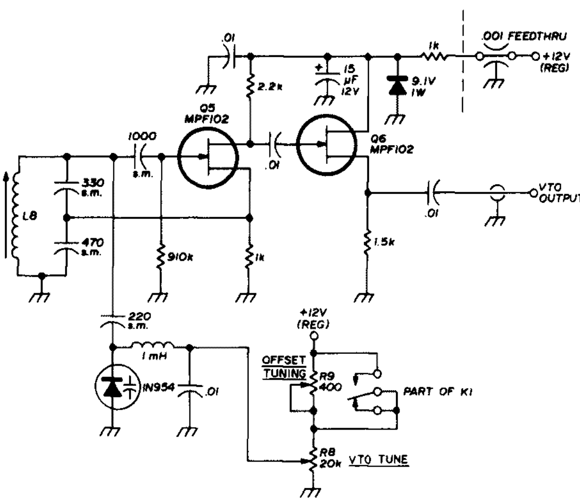


fig. 5. Varactor tuned oscillator (vto) is built in a separate enclosure which is attached to the main circuit board. L8 is 40 turns no. 32 on ¼" (6.5mm) slug-tuned form.

Remove the pilot lamp from L13 and place it across L15, coupling this stage to the receiver. Adjust R14 for about 5 to 10 mA of quiescent collector current at Q8. Then tune capacitor C6 for maximum signal strength in the receiver and maximum lamp brilliance.

To tune the 2N5993 linear amplifier, Q9, place a 10 watt, 50 ohm resistive dummy load across the antenna output, very loosely coupling this to the receiver and an oscilloscope. The easiest way to adjust the bias of Q9 is to monitor the modulation envelope on the oscilloscope while adjusting potentiometers R15 and R16 for the best ssb oscilloscope pattern and cleanest audio. In my transceiver, the quiescent collector current of Q9 was on the order of a few hundred milliamperes.

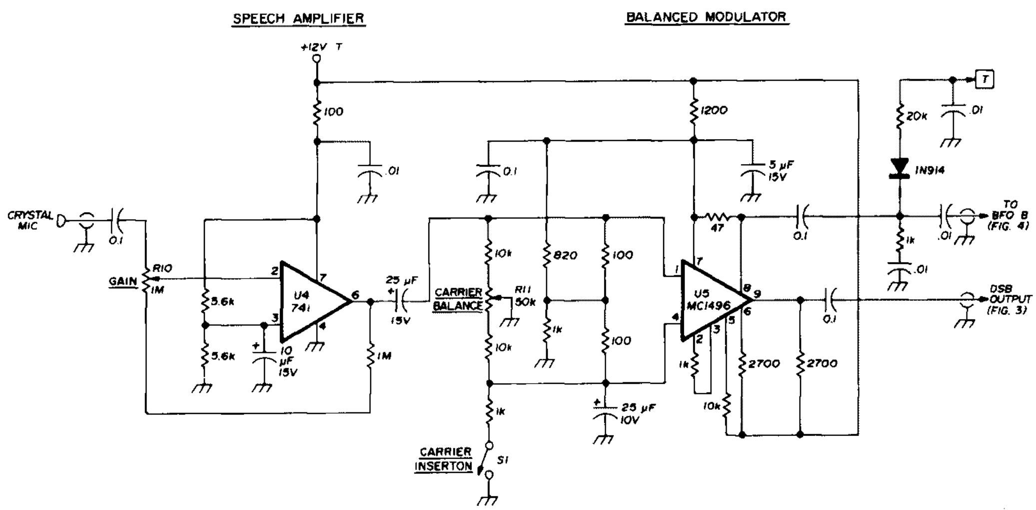
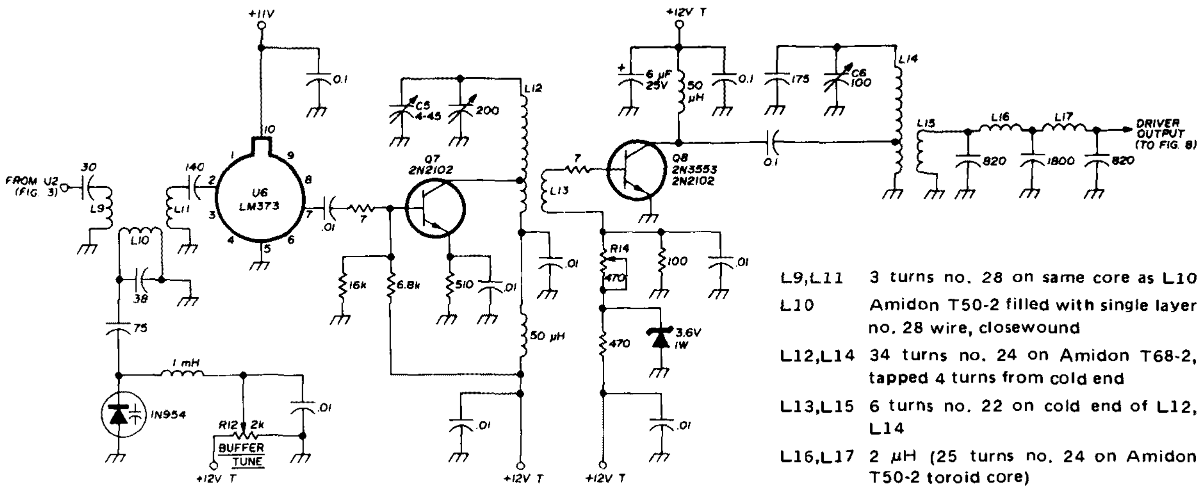


fig. 6. Speech amplifier and balanced modulator.



L9,L11	3 turns no. 28 on same core as L10
L10	Amidon T50-2 filled with single layer no. 28 wire, closewound
L12,L14	34 turns no. 24 on Amidon T68-2, tapped 4 turns from cold end
L13,L15	6 turns no. 22 on cold end of L12, L14
L16,L17	2 µH (25 turns no. 24 on Amidon T50-2 toroid core)

fig. 7. Buffer, pre-driver and driver stages.

The transceiver is easy to tune and operate, and it puts out a good ssb signal. Its modular design and construction make troubleshooting less frustrating and alignment simple. Although, with 7 to 8 watts output, this is really not a QRPp rig (according to the purists), it cannot compete with higher powered stations. However, by using good operating skill, tempered with patience, the operator will be rewarded with many contacts.

For those whose resources are unlimited, an rf power transistor specifically designed for linear ssb service could be substituted for the 2N5993 used at Q9. A 2N5992 would deliver about 10 watts PEP with a 12 volt supply, while a 2N5070 or a 40936 would deliver about 25 watts PEP using a 28 volt collector supply.

I wish to thank Charles Hill, W5BAA, whose article and personal communications helped in the formulation and design of this ssb transceiver.

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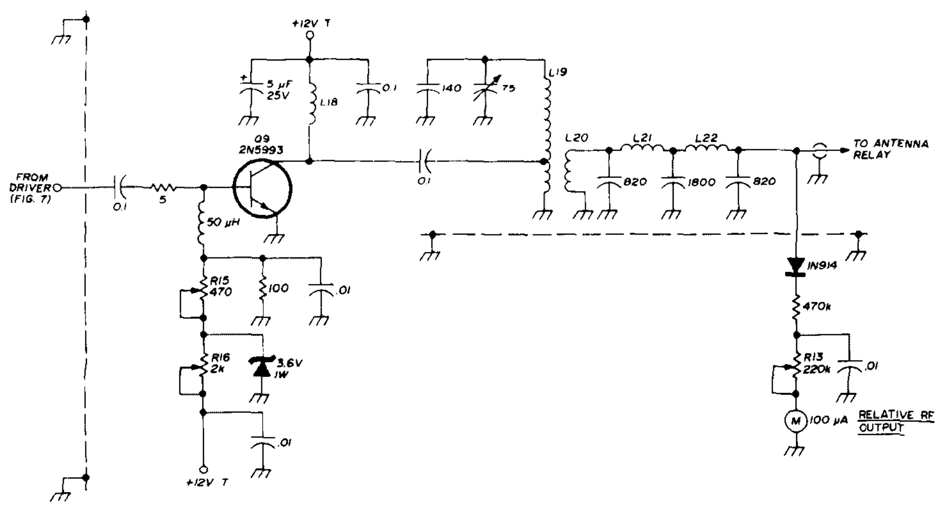
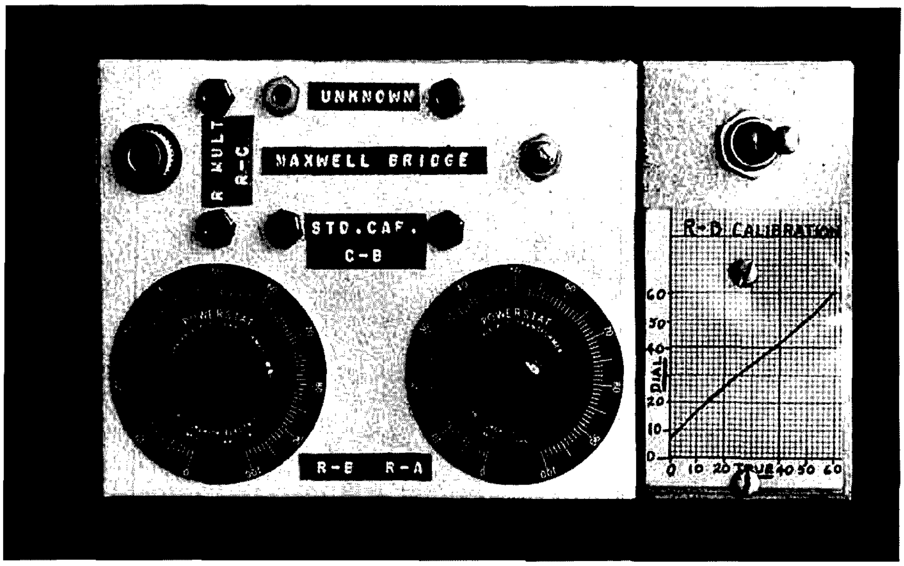


L18	2 layers no. 28, closewound on Amidon T37-2 toroid core	L20	6 turns no. 22 on cold end of L19
L19	34 turns no. 24 on Amidon T68-2	L21,L22	2 µH (18 turns no. 20 on Amidon T68-2 toroid core)

fig. 8. Final amplifier stage. The 2N5993 works well in this circuit, although not intended for linear service. Other transistor types may be substituted for increased output and linearity.



universal L, C, R bridge

Here's a useful instrument for test equipment that won't strain your imagination, credulity, or pocketbook. The idea isn't new; in fact it's of about the same era as the Wheatstone bridge. But for some reason, it seems to have been overlooked by the amateur fraternity. Known as the Maxwell bridge, it provides in one instrument the capability of measuring inductance, capacitance, and resistance.

Ohmmeters and capacitance meters are available in all price ranges. But not generally available (or familiar) to amateurs is a means of readily measuring inductance between 1 μH and 1 henry. The investigation of inductance bridges narrowed to the Maxwell bridge as by far the simplest, and also revealed its flexibility for resistance and capacitance measurements. The advantages of the Maxwell bridge are minimum number of components, calibration simplicity, measurement simplicity, driving-frequency independence, and measurement accuracy.

Most other bridges in common use have more than one reactive element as compared to one in the Maxwell bridge. Over a fixed range, R_a (see **fig. 1**) can be calibrated to read inductance values directly; and at a specified fixed driving frequency, R_c can be calibrated to read Q directly. In addition, R_a and R_b can be calibrated initially over their variable ranges by means of standard resistors. All measurements are direct products or quotients of direct readings. In the derivation of the basic equations for bridge balance, it is found that $2\pi f$ appears on both sides of the equation and cancels, with the result that bridge measurements are not affected by the driving source frequency. Finally, the nulls are unmistakable so that if the resistance arms are ±1%, the worst-case error would be approximately 3%, with the probability being 8 to 1 that the error would be less.

Referring to **fig. 1**, which is the setup for inductance measurement, and neglecting for the moment the potentiometer across the transformer, the equation for balance is $L_x = R_a \times R_c \times C_b$, where L is in millihenries, R in kilohms and C in microfarads. For the series resistance of L, $R_x = R_a \times R_c$ divided by R_b. Then $Q = 2\pi f L$ divided by R_x. Note that if L_x were vanishingly small, leaving R_x; and if C_b were removed, the bridge would be a straightforward Wheatstone type for resistance measurements with the same resistance balance equation as above. If a standard inductance is used in place of L_x, an unknown capacitor in the position of C_b can be measured. For this measurement, $C_x = L_{standard}$ divided by the product of R_a and R_c, with the units the same as for the inductance measurement. The series resistance of a capacitor is usually so small and the Q so large as not to be determinable with this bridge.

construction

All you need for layout and wiring is shown in the photo and the schematic. (The calibration curve in the photo was made because of nonlinearity in the low-resistance end of R_b in fig. 1.) I recommend a Bud 7x5x3 inch (17.8x12.7x7.6cm) Minibox box, which is about the smallest space into which you can squeeze the parts. Partition a 2-inch (5.1cm) space at one end to hold the signal-driving source, which consists of the

By J. H. Ellison, W6AOI, 1720 Holly Avenue, Menlo Park, California 94025

transformer, the IC and its components, and the switch and battery. The IC is a Signetics NE555 connected as an astable oscillator running at approximately 1000 Hz with the R and C values shown. The IC draws 6.5 mA. The battery is a 9-volt transistor battery. The transformer is a small audio type to provide dc isolation. Try to stay in a 3 to 1 ratio or less, but don't go down to the transistor interstage size. The pin jacks for the plug-in parts are Cinch-Jones with an all-Nylon body. For pins to match, find some plugs from any old Command set and saw out miniature banana-type pins by the handful.

The IC puts out a square wave at approximately 40% duty cycle, and you want some inductance to round off that square wave with its high harmonic content. Run the transformer secondary leads through a hole in the partition and connect them to the small pot, whose rotor is grounded. This arrangement is called a Wagner ground, a method for balancing stray internal capacitances to ground to get a perfect null. A first balance is made with R_a and R_b, then the Wagner ground is adjusted to deepen the null. Usually only one adjustment is required, then the R_a and R_b balance is perfect.

Bridge accuracy will depend on the linearity of R_a and R_b and the success you have in finding dial plates with graduations that match the angular range of your pots. If you can't match the two, the alternative is to make your own calibrated dial plates. The difficulty is that no uniformity exists among manufacturers in the angular range covered by the resistance winding. It goes without saying that the pots must be wire-wound to ensure reliable calibration and resetting. I found some surplus 1.75-inch-diameter (4.4cm) precision pots made by Technology Instruments, Acton, Massachusetts, and dial plates labeled "Powerstat" made by Superior Electric, Bristol, Connecticut, which matched exactly. You should be so fortunate!

Other manufacturers with models approximating these are Spectrol, Helipot Division of Beckman Instruments, CTS Corporation, Clarostat, and Mallory. Mallory has an inexpensive model designated MG, which might be satisfactory if you make your own calibrated dial plates. You'll find that the angular measurement of the resistance windings will range from 255 to 335 degrees, depending on construction. It's desirable, although not necessary, that the pots be metal enclosed to provide shielding. This shielding adds about 15 pF capacitance across the windings, which is equivalent to more than a megohm at the bridge driving frequency and may be neglected for all practical purposes. The phone jack should be insulated from the case, and for best sensitivity high-impedance phones should be used.

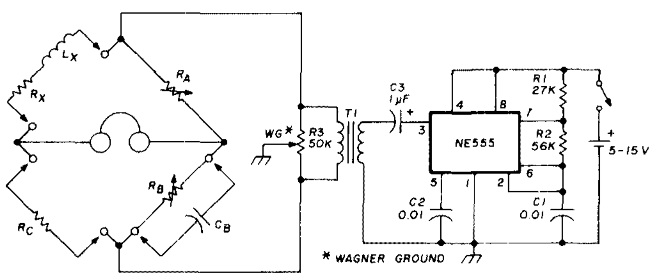


fig. 1. Maxwell bridge setup for inductance measurements. Wagner ground balances stray internal capacitances to ground to obtain a perfect null. Measurement ranges are shown in table 1.

R_a 1k wirewound linear pot
R_b 10k wirewound linear pot
T1 3:1 or less audio transformer

standards

You'll need resistance standards to plug into the R_c position and capacitance standards to plug in the C_b position. The minimum resistances will be 10, 100, and 1000 ohms, and minimum capacitances will be 0.01, 0.1, and 1.0 μF. Juggling the formulas will show that you can cover the same inductance range with different combinations of R_c and C_b, but the inductor series-resistance range will be different. For example, a C_b of 0.1 μF and an R_c of 100 ohms will cover the inductance range of 1.0 to 10 mH and a series resistance range of 10 to 100 ohms. A C_b of 1 μF and an R_c of 10 ohms will cover the same inductance range, but the resistance range will be 1 to 10 ohms.

table 1. Measurement ranges for the L,C,R bridge.

L (mH)	R (ohms)	C_b (μF)	R_c (ohms)
0.01 - 0.1	1 - 10	0.01	10
0.1 - 1.0	10 - 100	0.01	100
0.1 - 1.0	1 - 10	0.1	10
1.0 - 10.	10 - 100	0.1	100
1.0 - 10.	1 - 10	1.0	10
10. - 100.	100 - 1000	0.1	1000
10. - 100.	10 - 100	1.0	100
100. - 1000	100 - 1000	1.0	1000

If you can't get a distinct null, you have the wrong combination of C_b and R_c. You can work out other R and C combinations to expand the bridge range. You should make a table of combinations with the corresponding L_x and R_x ranges.

Using the plug-in method for C_b and R_c provides the greatest versatility and most compact bridge size. If you used internal switching systems, you'd add considerable mechanical complexity and stray capacitive coupling. Some of the components are external to the enclosure and some are within, which adds to mutual shielding. No proximity hand effects should occur when balancing the bridge.

The standards mentioned will cover inductances between 10 μH and 1 henry, with corresponding resistance ranges between 1 and 1000 ohms. For capacitance measurements between 10 pF and 10 μF, you'll need a 1 mH standard and two additional resistance standards, which you can calculate. These same resistance standards will be useful in making resistance measurements, which can be determined with much greater accuracy than with an ohmmeter.

A final word: an unavoidable interaction exists between the reactance and resistance balances. Care must be taken to ensure a *complete* null to avoid measurement errors. If the approximate R_x is known beforehand, correct measurements are expedited.

ham radio

repair bench

Michael James

troubleshooting by resistance measurement

In addition to the signal tracing, signal injection and voltage measurement techniques which I have discussed in previous columns, troubleshooting by resistance measurement in another basic technique which can be used to pinpoint circuit problems. One of the best arguments for resistance troubleshooting, of course, is safety. You can check out many of the stages in your transmitter, for example, without any lethal voltages applied, and when you must work on high-voltage power supplies resistance troubleshooting is the *only* method which is recommended.

A few amateurs will contend that resistance measurements don't give the complete story because some components in a high-voltage circuit that check out okay by resistance measurement may break down under stress. That's true, but you can usually spot these problems by a visual inspection – they smoke, or arc. Some short circuits occur in transmitters only after high voltage and drive have been applied, so resistance or voltage measurements are of little use, but once again, you can usually *safely* track down the difficulty by a visual inspection and the process of elimination.

Resistance troubleshooting can be done with a simple volt-ohm-milliammeter or vom, and most amateurs have one. In addition to being relatively inexpensive, rugged, and simple to operate,the volt-ohmmeter doesn't have to be plugged into the power line so it's completely portable. Some of the new solid-state electronic volt-ohmmeters are also portable, but they're considerably more expensive and can sometimes give erroneous readings when used in operating rf equipment.

There are any number of good quality volt-ohm-milliammeters on the market which are suitable for amateur use, ranging from miniature, imported models that sell for less than $10 to more accurate instruments which go for $100 or more. Although many of the pocket-size meters are good investments, if most of your troubleshooting is done on the workbench, you'd probably be better off with a larger vom with an easy-to-read 4 or 5 inch (10-13cm) scale. Insist on dc accuracy of at least 2%, and don't be misled by the fact that 1% resistors are used in its construction. The use of 1% resistors doesn't necessarily mean that the readout is very accurate – the meter movement itself may be non-linear.

Other features to look for are 20,000 ohms/volt sensitivity (minimum) and a meter protection circuit. Some meters are completely burnout proof, and you pay extra for this premium, but it might save you money in the long run.The meter shown in the photographs, a Triplett model 60, is both burnout proof and shockproof, so I should never have to replace it. (It once got knocked off my bench on to the concrete floor and sustained absolutely no damage.)

When selecting a vom the most important factor is accuracy. Dependable ohmmeter measurements are not much use if you can't read them accurately, and you can't begin to interpret their meaning unless the measurements are accurate. Following are four suggestions which will help you to obtain accurate measurements with your ohmmeter:

1. Calibrate the instrument before using it. Place the vom in the position you're going to use it, standing upright or lying down, with the function switch in the *off* or *dc voltage* position, and with the test leads plugged in but not shorted together, make sure the meter needle rests exactly on the zero mark at the left edge of the scale. You may have to adjust the mechanical zeroing screw that is just below the meter face (see fig. 1). If it's been awhile since you've made the zero adjustment you

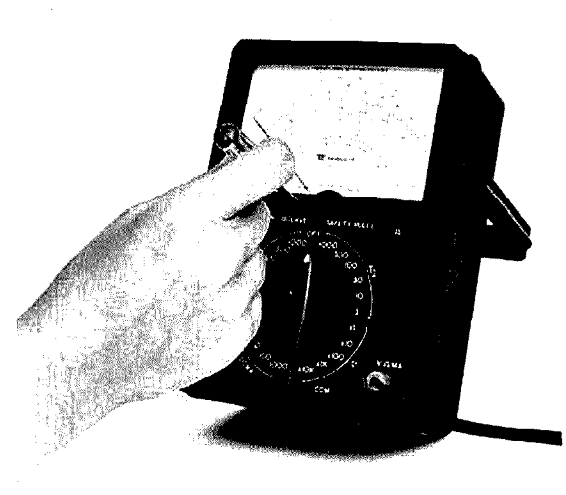


fig. 1. Before making any resistance or voltage measurements, make sure the meter needle is zeroed. If it's not, adjust the mechanical zeroing screw just below the meter face on the front of the vom. This adjustment shouldn't be required very often, but you should always check the mechanical zeroing before you use the instrument.

fig. 2. The needle should move up scale to the zero ohms mark when the test leads are shorted together. If the needle doesn't align with the zero ohms mark, adjust the "ohms" or "zero ohms" control on the front of the instrument.

fig. 3. The meter pointer should rest somewhere near midscale for best accuracy when making resistance measurements. This can be accomplished with the "range" switch which is calibrated in decade increments.

may have to lightly tap the meter glass to jar loose any needle friction.

2. Zero ohms adjust. Turn the function switch to the *ohms* position and short the two test leads together. The needle should move up to full scale (since the ohms scale reads from right to left, zero ohms is at full scale). If the pointer doesn't line up with the zero ohms marker, adjust the *zero ohms* control on the front panel of the instrument (see fig. 2).

3. Range switch. For accurate resistance measurements the range switch should be set so the meter operates in the upper two-thirds of the resistance scale. This is because the resistance scale is non-linear and the scale on the left edge is compressed, making it difficult to read accurately. Moving the range switch up one or two positions will move the needle toward the right into a portion of the scale where the calibration marks are further apart.

4. Properly interpolate the scale reading. Most errors in resistance measurements are due to simple technician error in properly multiplying the ohms scale reading by the *multiplier* indicated by the range switch. It's pretty easy to drop a zero (or add one) and end up with the wrong measurement. When the pointer is on the left side of the scale you'll have zeroes in the scale reading as well as in the multiplier, so be particularly careful and use scratch paper if you're unsure.

series resistance circuits

Before you can troubleshoot an electronic circuit with resistance measurements you have to recognize the different kinds of resistance paths which you may find in an actual circuit. There are dozens of resistors in practically every electronic circuit and, in addition to resistors, there are a number of other components which will show readings on an ohmmeter such as transformer windings, electrolytic capacitor leakage, forward and backward resistance of semiconductor diodes, vacuum tube filaments and, of course, transistors.

One of the tricks of resistance troubleshooting is finding resistance paths that aren't where they should be, or paths that have too little (or too much) resistance. To do this you'll have to learn how to spot the resistance paths on a schematic diagram, and how to figure out where and what they should be at different points in the chassis wiring.

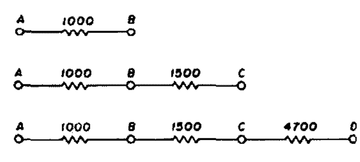


fig. 4. When resistances are in series their values add directly.

Simple series resistance paths, such as those shown in fig. 4, are both the easiest to spot and the simplest to analyze. It would be difficult to mistake the A-B path shown first – it's 1000 ohms between points A and B. In the second series path the 1000 ohms between A-B adds to the 1500 ohms between points B-C for a total resistance of 2500 ohms. The third series path, between points A-D, is also pretty easy, adding up to 7200 ohms.

It's important to recognize in the simple series circuits of fig. 4 that adding the B-C and C-D resistance paths had no effect on the A-B resistance. In all three

cases an ohmmeter connected to points A and B would indicate 1000 ohms. You can add as many series resistors as you want and A-B will remain at 1000 ohms.

Another important point concerning series resistances is that you can measure each of the individual paths independently (B-C or C-D, for example) and the other series paths will not interfere with the reading. You can also directly measure path A-C if you wish (2500 ohms), or a path A-D (7200 ohms).

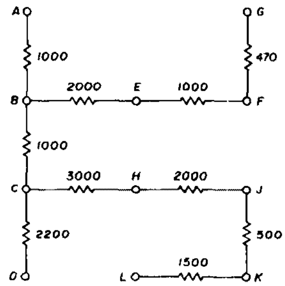


fig. 5. These resistances, although not arranged in straight lines as those in fig. 4, are in series and can be added directly.

Resistance paths don't necessarily have to be drawn in a straight line to be in series. An example is shown in **fig. 5**. An ohmmeter connected between points A and D measures the A-D path directly; the resistance is 4200 ohms. None of the other resistances in the circuit has any effect on path A-D because they are not in series with it (or shunted across it). The same thing is true about path A-L. In this series circuit only the resistance between points A and L have a bearing on the resistance of the path – it can be measured directly by simply connecting the ohmmeter leads to points A and L (9000 ohms). The resistances in paths B-G and C-D are ignored because they're not in series with path A-L. If you're in doubt, trace path A-L: A to B, to C, to H, to J, to K, to L.

Now consider the path from F to K (9500 ohms). It goes from F to E, to B, to C, to H, to J, to K. An ohmmeter connected between F and K measures only that path, and indicates 9500 ohms. None of the other paths in the circuit have any effect because they are not in series with path F-K.

Suppose you were troubleshooting a circuit like **fig. 5** and measured 8000 ohms between B and J. The schematic shows series resistances totaling only 6000 ohms in the B-J path. At least one of the resistors has changed value and you have to determine which one it is. One way to find the answer is to individually measure each resistance. Since there are several resistors in the circuit, that can take a fair amount of time. A better way is to work your way through the circuit from one end to the other. Leave one ohmmeter lead on B and move the other lead from J to H. If the reading drops to 4000 ohms it means that path H-J is at fault. If the reading only drops to 6000 ohms, however, it means that H-J is okay and the trouble is in the B-H path. One more measurement should locate the bad resistor.

Those are the basic principles of troubleshooting around series resistance circuits. There are plenty of series resistance circuits in electronic equipment, and as long as the resistors are in series, tracking down trouble is pretty easy. If the resistors are in parallel, however, as will be discussed next, the task is a bit more difficult – at times it can be downright confusing.

parallel resistors

Three very simple parallel resistance paths are shown in **fig. 6**, but you're not likely to find anything as simple as this in any electronic equipment. Nevertheless these simple circuits are a good starting point. So far as an ohmmeter is concerned, the path between points A and B is merely path A-B. As you can see from the diagram, however, it isn't nearly that simple because there are actually two resistance paths in the first two circuits (A-B and C-D) and five actual resistance paths in the third.

The resistance of path A-B is 500 ohms because the resistance of two parallel resistances of the same value is half the resistance of either resistance by itself, (similarly, the parallel resistance of three equal-value resistors is one-third the value of one resistance, and the parallel resistance of four equal-value resistors is one-fourth the

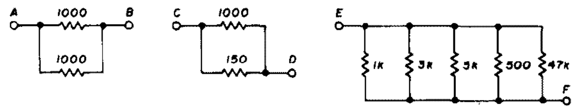


fig. 6. Parallel resistance paths are more difficult to calculate. A formula is given in the text below for calculating the equivalent resistance of a parallel resistance network.

value of one resistance, etc.). This is another way of saying that resistances in parallel add inversely. A general equation which can be used to calculate the parallel resistance of any number of resistors, of any value, is

$$\frac{1}{R_T} = \frac{1}{R1} + \frac{1}{R2} + \frac{1}{R3} + \frac{1}{R4} + \frac{1}{R5} + \cdots \frac{1}{R_n} \quad (1)$$

where R_T is the parallel resistance and $R1$, $R2$, $R3$, etc. are the values of the resistors in each of the parallel branches. If only two resistors are in parallel, **eq. 1** can be rewritten into the form

$$R_T = \frac{R1 \cdot R2}{R1 + R2}$$

Although it may not be immediately obvious by looking at the parallel resistance formulas, whenever a resistance is added in parallel to a circuit, it always lowers the resistance between two points. In fact, the parallel resistance is always less than the lowest resistance in any branch of the parallel path. The C-D path in **fig. 6** shows a situation where one branch of the parallel circuit has much lower resistance than the other. Calculated by the parallel resistance formula, path C-D has a resistance of 130.4 ohms. This would be the resistance measured between points C and D with an ohmmeter.

If you should run into this circuit in a bench test, consider what it means if you measure path C-D and find

a reading of 150 ohms. You can assume the 1000-ohm resistance has opened up or has changed value to some extremely high resistance (in general, if a *parallel* resistance is 100 times higher than the other resistor in the circuit, it will lower the parallel resistance by approximately 1%). Conversely, if you check path C-D with your ohmmeter and find that it measures 1000 ohms, the 150-ohm resistor is at fault.

To check the use of the parallel resistance formula, calculate the resistance between points E and F in **fig. 6.** Your answer should be 281.32 ohms. If you remove the 47k resistance and run through the calculation again, you'll find the parallel resistance of the circuit has increased only slightly, to 283.02 ohms. This is because the 47k resistance is nearly 100 times greater than the smallest resistance in the circuit.

One of the problems with parallel resistance paths in electronic equipment is that they are seldom as obvious as those in **fig. 6.** In practice parallel resistance circuits may have branches all over the chassis, and they are often arranged in rather strange shapes as shown in fig. 7. At first glance, for example, path A-B looks like a

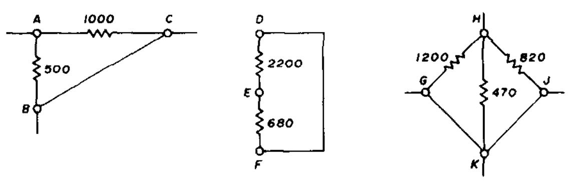


fig. 7. Although it might not be readily apparent, these three circuits are actually parallel resistance circuits. In practical circuit work there may often be hidden parallel circuits which can fool the unwary.

simple 500-ohm path. Since B is connected to C, however, resistance A-B is actually in parallel with resistance A-C for a total parallel resistance of 333 ohms. Connect an ohmmeter between points A and B (or between A and C) and this is what it would indicate.

Now consider the D-E-F resistance path in **fig. 7.** Although the two resistors appear in series, they're really not by virtue of the fact that D is connected directly to F. Any current which enters the circuit at D flows through both resistors to reach E. An ohmmeter connected at points D and E measures the combined parallel resistance of the two and indicates 519 ohms.

Now that you're looking for sneak parallel paths it should be easy to sort out the parallel resistors in path H-K. Points G, J and K are electrically the same so the three resistors are in parallel, and an ohmmeter connected between points H and K would measure 239 ohms.

These simple parallel resistance hookups are illustrated primarily so you won't overlook a parallel resistance connection just because it isn't obvious. In many circuits, in fact, many of the hidden parallel resistance paths are through other types of components and don't actually appear on the schematic as resistance paths. In every case, however, any parallel resistance path always *lowers* the ohmmeter reading to a value less than the lowest value in any branch of the parallel path. If you forget that one simple fact, resistance troubleshooting can be very confusing – remember it, and you can use this technique to pin down some very elusive faults.

series-parallel resistors

In most electronic circuits you won't find only series or parallel resistance, but a combination of the two. A few of the possibilities are shown in **fig. 8.** In the first

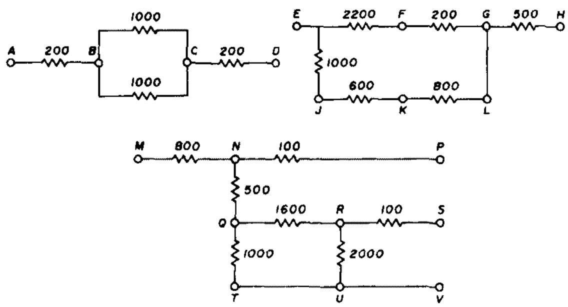


fig. 8. These series-parallel resistance networks are more like those which you will find in real circuits. To figure out the total path resistance first calculate the equivalent parallel resistance, then add the series resistance. Examples are given in the text.

circuit paths A-B and C-D are in series with the two-branch *parallel path,* B-C. To find out what resistance to expect if you connected your ohmmeter between points A and D, first calculate the resistance of path B-C; it's 500 ohms. Therefore, the total resistance from A-D is 200 + 500 + 200, a total of 900 ohms.

The second series-parallel circuit in **fig. 8** is slightly more complicated, but it's not difficult to figure out if you look at it carefully. The path from E to G is made up of two parallel branches, both of which have more than one series resistance. To calculate the parallel resistance of path E-G, first add the series resistance in each of the parallel paths E-F-G and E-J-K-L-G, calculate the parallel resistance, and then add the series path G-H:

path E-F-G = 2400 ohms
path E-J-K-L-G = 2400 ohms
path E-G = 1200 ohms
path E-G-H = 1700 ohms

The third series-parallel circuit in **fig. 8** looks even more complex, but if you study it carefully for a moment you'll see that it's quite easy to figure out the total resistance between any of the points of the circuit. Path M-P, for example, is a simple series path equalling 900 ohms. None of the other resistors in the circuit have any effect because they're not in series or parallel with that path. Path M-S, however, has a parallel branch which must be considered (Q-R is in parallel with Q-T-R); 1600 ohms in parallel with 3000 ohms or 1043 ohms parallel resistance. The total resistance between points M and S is

800 + 500 + 1043 + 100 = 2443 ohms

This is what your ohmmeter would measure if it were connected across M and S.

There is one other resistance path to consider in this circuit: M to V. In this circuit path Q-R-U is in parallel with path Q-T-U (3600 ohms in parallel with 1000), for equivalent parallel resistance of 783 ohms. Therefore, the total series resistance between M-V is

$$800 + 500 + 783 = 2083 \text{ ohms}$$

hidden resistance paths

There are two components which will most likely give you false ohmmeter readings: semiconductors and electrolytic capacitors. Since electrolytics are used for power supply filters and decoupling, they are spread throughout any electronic circuit, so it's pretty hard to avoid them and their dc leakage currents which register as resistance on your ohmmeter. In solid-state equipment many of the coupling capacitors are also electrolytic types (as opposed to paper or plastic capacitors used in vacuum-tube circuits), and these can give you a lot of grief if you don't know what you're looking for. Bipolar transistors also give problems because each of the junctions looks like a diode to your ohmmeter, with greater resistance in one direction than in the other.

Fortunately capacitance leakage paths and semiconductor resistance paths follow certain patterns. If you know the patterns you won't be fooled -- and even if you are fooled at first it shouldn't take you too long to get back on the right track. Here are a few of the patterns to keep in mind:

1. **Diode** action can usually be checked by simply reversing the ohmmeter leads. Consider the circuit of **fig. 9**. Suppose the resistance reading from the 12 Vdc terminal to ground (power *off*, of course) measures about 150 ohms. It looks like filter capacitor C1 has shorted. To check, move the test lead to the junction of R1-C1 -- a very low resistance there seems to confirm that capacitor C1 is shorted. Before you jump to conclusions, however, reverse the test leads and repeat the measurement. The low reading will probably disappear. Why? Because the internal battery of the ohmmeter has forward biased the rectifier diodes, causing them to provide a low-resistance path to ground through the transformer center tap.

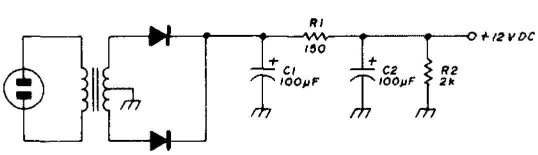


fig. 9. Simple solid-state power supply. Measured resistance from the 12 Vdc terminal to ground may be lower than expected because of a hidden path to ground through forward-biased rectifiers.

Whenever the negative side of the ohmmeter battery (sometimes the black test lead, but not always) is connected to the cathode end of a semiconductor diode and the positive lead is connected to the anode, the ohmmeter reads the diode's forward resistance (usually less than a few ohms). Reversing the test leads reverse biases the diode and measures the diode's back resistance which is typically 100k or more. Therefore, if you find a resistance path which is much lower than it apparently should be, reverse the test leads to make sure a semiconductor diode (or bipolar transistor) isn't causing the lowered resistance reading.

2. **Transistor leakage** can cause the same sort of measuring problems as diodes because both the base-collector and base-emitter junctions are, in essence, diodes so a transistor junction that is forward biased by the ohmmeter battery looks like a low-resistance path. Consider the circuit of **fig. 10** where an npn transistor is used in a typical rf mixer circuit. Suppose you connect your ohmmeter across resistor R2 to measure it. Instead of the expected 2700 ohms the meter reads about 50 ohms and you figure R2 has changed value. If you reverse the meter leads, however, you measure about 2100 ohms.

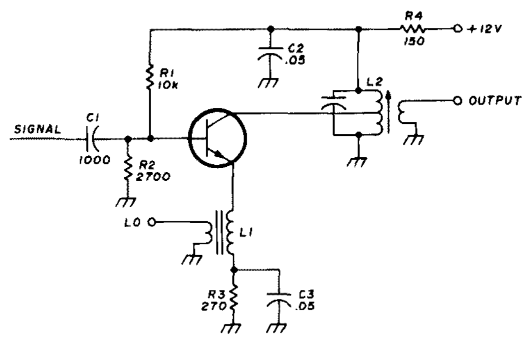


fig. 10. Rf mixer circuit using a bipolar transistor. There are many hidden resistance paths in this circuit which can be confusing if you're not aware of them.

The reason the ohmmeter indicated 50 ohms in the first case, of course, is because the base-emitter and base-collector junctions were forward biased, so the base-emitter junction resistance (about 50 ohms) provided a parallel path to ground through the 270-ohm emitter resistor, R3, and the forward-biased base-collector junction provided a path to ground through L2. When the test leads are reversed the two transistor junctions are reverse biased and your ohmmeter measures the parallel resistance path provided by R1 *and* R2 (about 2125 ohms).

When checking npn transistor circuits remember that the base-emitter (or base-collector) junction is forward biased when the base is positive and the emitter (collector) is negative. In pnp transistor circuits the two junctions are forward biased when the base is negative and the emitter (collector) is positive.

3. **Electrolytic capacitors** have leakage currents which the ohmmeter reads as resistance. The leakage resistance of most electrolytics is in the range of 50 kilohms or so, so learn to allow for it when you're checking resistances along power supply lines. One clue to hidden resistance circuits which are caused by electrolytic capacitors is

that the resistance reading tends to increase if the test prod remains on the point. This is because the ohmmeter battery is charging the capacitor, and as the capacitor nears full charge it draws less current, which the ohmmeter interprets as a higher resistance. Reversing the test leads will cause most electrolytics to instantaneously measure as short circuits, but if the leads are left in place the capacitor will once again begin to charge and the meter pointer will move up scale to 50 kilohms or so, the normal leakage resistance of the capacitor.

Hidden resistance paths which are drawn elsewhere on the schematic can also cause erroneous resistance measurements. If you were to measure the resistance from the plate of V1 to ground in **fig. 11** you might expect to find an infinite reading. Instead you find a low resistance reading caused by the voltage-dividing network in the screen circuit of V12 which is located on the other end of the schematic (**fig. 11B**). Hidden circuits such as this

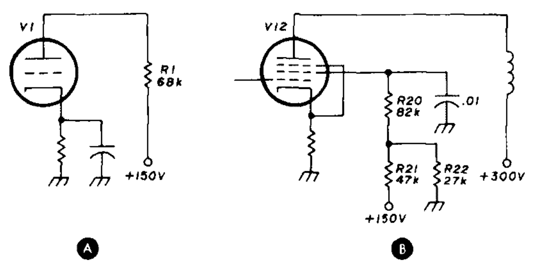


fig. 11. You might expect a resistance measurement from the plate of V1 to ground to be 100k. Actual measurement is much lower because of voltage-dividing screen network for V12 — a parallel resistance path which is hidden because of the way the schematic is drawn.

are usually a bigger problem in vacuum-tube equipment than they are in solid-state gear because base-bias networks are common practice in transistor circuits and you'll learn to work around them. Bias networks are less common in tube circuits, however, but when they do occur you'll probably find them at some remote point on the schematic where they're easily overlooked.

These are the four most common hidden resistance paths which can give you erroneous readings when you're troubleshooting by resistance measurements. When your ohmmeter readings are different from what the schematic leads you to expect, here are some quick checks to pinpoint the cause.

1. If a reading is low, reverse the ohmmeter leads. If the reading increases, there's diode action in the circuit which can be traced to a diode, transistor or a power supply rectifier. Only slight differences in readings when you reverse the test leads can usually be traced to electrolytic capacitors which don't form under reverse polarity.

2. If an ohmmeter reading starts at the low end of the scale and builds up slowly, an electrolytic capacitor is causing it through normal charging action. Reversing the test leads gives an even lower reading (often off scale) which builds back up again to the higher one. The final reading should be 50 kilohms or greater and depends upon the condition of the capacitor. The clue here is the charging action.

3. If you're still looking for the cause of a low resistance reading, have eliminated semiconductors and electrolytic capacitors, and can find no hidden parallel paths at other points on the schematic, you're probably on the trail of the circuit problem.

Sometimes it's faster and easier to eliminate the effects of parallel resistance paths than it is to get around them. One troubleshooting trick that often helps is not to make resistance measurements with respect to ground — don't clip the common lead of the ohmmeter to the chassis. Make each resistance measurement between points *in the circuit*. This won't eliminate all the hidden parallel circuits, but it will isolate some of them, which is a help.

If necessary, disconnect unwanted parallel resistance paths when they get in your way. If you carefully plan your point-to-point resistance measurements, this shouldn't be required very often, but when it does happen it's usually a simple matter to temporarily unsolder the parallel component or resistance branch which is giving you fits.

On printed-circuit boards you can often unsolder one lead. Another technique the professionals use is to cut a slit across the copper foil with an Xacto knife to disconnect the offending circuit branch. The circuit is easily restored by a solder bridge across the narrow slit.

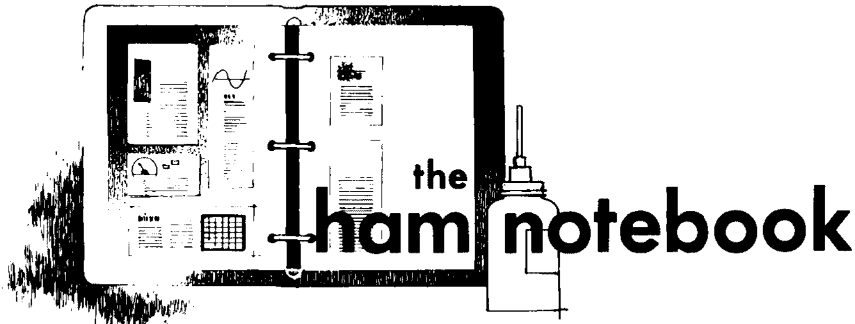
You can sometimes eliminate parallel circuit paths by pulling a tube or a transistor out of its socket. However, with fewer and fewer sockets on modern printed-circuit boards, this isn't always possible. In those cases it's easier to cut a slit in the circuit trace.

As a *final* resort, when you can't analyze a circuit any other way, you may have to disconnect and test each component in the circuit one at a time. This is a time consuming process, but if you've done your premiminary testing carefully, and eliminated most of the parallel resistance paths by one of the previous techniques, you'll have the problem narrowed down to a small section of the circuit so there shouldn't be too many parts to test individually.

conclusion

Although most manufacturers provide resistance charts in their instruction manuals, in those cases where charts are not available resistance troubleshooting is based entirely on figuring out from the schematic diagram just what a particular resistance path should be, and then measuring it with your ohmmeter. If the resistance isn't what you expect it to be you must isolate the one that's incorrect. In the coming months I will show how the technique of resistance troubleshooting, when coupled with voltage measurements and signal tracing, can be a very valuable tool for locating circuit problems in amateur equipment.

ham radio



Mini-Mitter II modifications

The Mini-Mitter II* is no longer available in kit form. American States Electronics is now building an assembled MM-2C for sale primarily to government agencies for approximately $600.00. I was able to pick up a used unit with a broken whip antenna. The following is offered on replacing the antenna together with some data on replacements for in-house brand ICs and transistors.

The stock antenna is 65 inches (1.65m) long. The manufacturer reports that his supplier no longer makes this length rod and has substituted a smaller one, as found in CB sets. No replacements are available. Lafayette Radio in their latest catalog lists a 67½-inch (1.7m) antenna (stock no. 99F32070), which is the only close replacement I've found. The base diameters differ, but replacement can be accomplished as follows:

1. Cut off the heat-shrink tubing covering the loading coil. Be careful not to nick any turns of the coil.

2. Two screws will be exposed, which hold the antenna rod and collar. Remove these two screws and pull off antenna and collar.

3. The collar is soldered to the antenna rod. Clamp the rod in a vise, heat the collar with a torch, and remove.

4. To accommodate the Lafayette rod, the collar must be bored out to 0.375 inch (1cm). Resolder the new antenna to the bored-out collar and reassemble to the loading coil.

5. Obtain a piece of 1½-inch (3.8cm) heat-shrink tubing to fit over the coil. Finding tubing this large may be the hardest part of the whole job.

*Ken Pierce, W6SLQ, "Mini-Mitter II," *ham radio*, December, 1971, page 72.

Equivalents for the Mini-Mitter house-brand ICs and transistors are listed below:

FE5245	2N4416
A158C	HEP737
ASE77	TRW PT2677B
ASE400	RCA 3020A
ASE401	RCA 3021
ASE408	RCA 3028A

ASE's latest transceiver is the model MM-2C. Apparently there have been some modifications to decrease standby current drain in the audio output circuit and to increase transmitter output power. Perhaps a reader with a newer unit might comment on this.

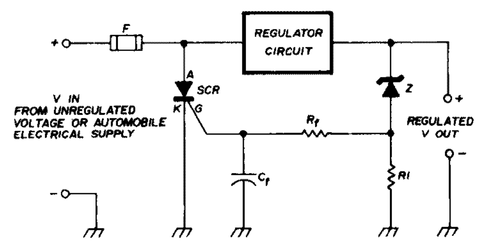
Charles King, K1ETU

overvoltage protection

Most solid-state electronic equipment in use today specifies an input voltage of 13.8 Vdc ±15%. This represents a range of 11.475 to 15.525 Vdc. If the series element in the regulated power supply short circuits, the entire output from the rectifier (typically 18 to 30 Vdc) will appear at the input to your equipment. Few semiconductor circuits in popular use can withstand such abuse.

The crowbar (overvoltage protection circuit) shown in fig. 1 can be added easily to your power supply or automobile electrical system for just a few dollars. It will detect any overvoltage condition and immediately shut down the supply voltage before it fries your equipment.

Operation is simple. When V_{out} reaches a set value (determined by Z), the scr gate triggers, conducting heavily, and the fuse blows. This circuit has the advantage of not loading the regulator should the scr fire because of some transient condition. More important, the fuse immediately opens because of the short circuit condition. This fast-turn-off feature is important, because the time required to open a fuse is a function of the load current, fig. 2. A typical 10-watt transceiver may draw 2 to 3 amps on



F	fast-blow fuse, typically 150% of nominal equipment load current
R1	1000 ohms typical. Provides leakage current for zener diode
Rf, Cf	lowpass filter to prevent noise spikes from turning on SCR. Rf - 100-1000 ohms typical, Cf = 0.10-0.05 μF typical
SCR	rating determined by voltage and short-circuit current requirements
Z	zener diode voltage as required to detect over-voltage condition

fig. 1. Overvoltage protection (crowbar) circuit for regulated power supplies.

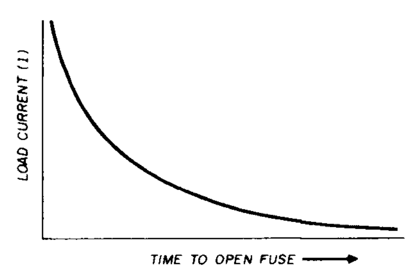


fig. 2. Relationship of load current and fuse rupture.

transmit but only a tenth or so of this value in receive, permitting a potentially damaging situation.

To check the crowbar, temporarily install a 500-ohm resistor in the scr anode circuit (fig. 3). Increase output voltage to the maximum desired and check to see if the scr fires properly, which will be indicated by a sudden voltage increase as measured across the 500-ohm resistor.

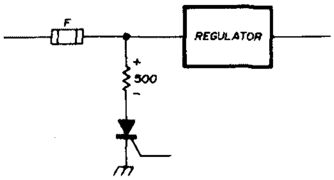


fig. 3. Test circuit for checking crowbar.

If necessary change Z or add diodes in series (germanium 0.3 V or silicon 0.6 V) to obtain the desired trip voltage. When all is well, remove the resistor. The scr will conduct until power is removed from the circuit. This circuit is analagous to buying insurance: you may never use it, but you may be sorry someday without it.

Ed Pacyna, W1AAZ

R-392 receiver mods

The R-392 receiver is hot, stable, and provides full coverage between 0.5 and 30 MHz. However, the compromises involved in modifying it for 28 Vdc on filaments and plates need shaping up. Here are some suggestions.

First bring out the plate and filament leads separately. The receiver is happier with somewhat more than 24 volts on the plates, somewhere between 30 and 35 volts. Build a supply that will provide over 30V from a bridge rectifier. Filter the supply with 3000 μF or more (the receiver tolerates 2 to 4 V of hum). Power the entire receiver this way first. The tubes in my set withstood 37 V for a few hours. Next add a dropping resistor (2 ohms, 20 watts) to the filament lead. Leave the plate lead on the higher voltage.

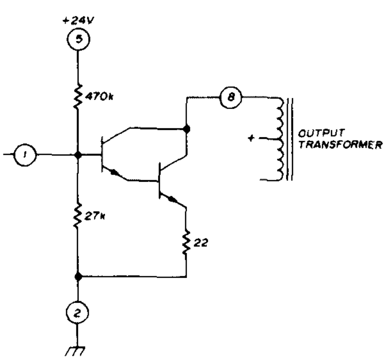


fig. 4. Simple audio replacement for V608.

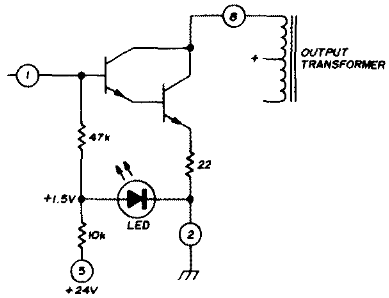


fig. 5. Improved audio replacement.

The audio output tube has to go. This tube pulls a total of 1.3 watts, plate and filament. The designers must have known it had to go: the manual shows a plug-in transistor substitute.

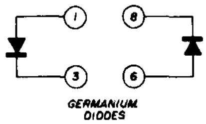


fig. 6. V602 and V603 replacements.

However, their substitute uses 15 components including four transistors and two transformers. A better substitute is shown in fig. 4. Use a single Darlington power transistor with no extra heatsink; connect the collector to existing pin 8, emitter through 22 ohms to existing pin 2 and ground, and base to existing pin 1. Bias the Darlington with 27k to pin 2 and ground and 470k to pin 5 (+28 V, formerly the screen supply). Build this assembly on an octal plug. Adjust the 470k resistor to obtain 0.2 to 0.7 V across the 22-ohm resistor.

An improved circuit is shown in fig. 5. A LED makes an excellent 1.5 V zener — much better than a zener at low current. I used 10k to a LED to obtain 1.5 V and 47k from the LED to the Darlington base. That's a total of five components. The dropping resistor in the power supply can now be increased to 4 ohms. You have removed 16 watts from the filament string.

Next we operate on the detectors. V602, V603 are 12AU7s with filaments in series. They are used as four diodes for detectors and agc. Pull out V603 and insert two germanium diodes into the socket holes, one with anode to pin 1 and cathode to pin 3 (see fig. 6), and the other with anode to pin 6 and cathode to pin 8. The set should work, even with V602 dead. Insert two more diodes in the same places on V602 socket to get the squelch rectifier and agc working again.

You can now increase the power-supply dropping resistor to about 8 ohms, and the filament string will have decreased from 3 to just over 2 amps. The radio has now become practical, and you haven't had to take the chassis apart. The next step should involve fet substitutes, a very practical possibility with only 30 volts on the plate bus.

N. J. Thompson, KH6FOX

selecting white noise diodes

Several articles have appeared in *ham radio* on how to build an rf noise bridge for measuring antenna impedance. The following method for finding a suitable noise diode is the utmost in simplicity.

The equipment needed is a variable voltage supply between 5 and 10 volts, a variable resistor box or potentiometer, and your communications receiver. If you don't have a variable power supply, use a 9-volt battery. However, if you wish to know more about the characteristics of each diode, a variable supply will be needed and each diode can be catalogued. The test setup is shown in fig. 7.

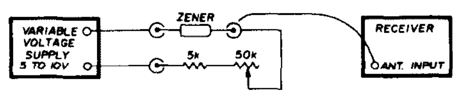


fig. 7. Setup for testing "noise generator" zener diodes.

Turn on your communications receiver and run a short antenna wire to within a few inches of the diode under test, apply 9 volts to the diode with a series resistor of about 10k to 40k. If no noise is heard, reverse the diode polarity. When the noise has been maximized for the range of 7½ to 9 volts, use

an ohmmeter to measure the total resistance of the variable resistor plus the fixed resistor. About ten per cent of the diodes tested gave an output sufficient to produce a roar from the receiver loudspeaker, and the optimum resistance was 33k.

With a variable power supply you can optimize and catalog each diode for voltage and resistance. I selected my diodes for the noise bridge using 9 to 7½ volts at some given resistance, which provides the greatest use of a 9-volt battery.

Lloyd Jones, W6DOB

goral oscillator notes

The Goral crystal oscillator circuit described by Don Stoner in *ham radio** appears to be excellent in many respects. I have found, however, that the proper value of C2 in **fig.** 4 of the original article is a critical function of the capacitance for which the crystal is calibrated. Crystals for the GE Progress Line, for example, are ground to operate into a 10-pF load and will not oscillate on their proper frequency using 20 pF as the value of C2. Data on two different crystals for a GE Progress Line receiver are shown in **fig.** 8. A value of 12 pF for C2 is more suitable as it allows the crystal to be netted using an 8-pF trimmer capacitor at C1. The data also illustrate the wide frequency range over which the oscillator will operate when different values of C1 and C2 are used.

Robert E. Cowan, K5QIN

*Donald L. Stoner, W6TNS, "High-Stability Crystal Oscillator," *ham radio*, October, 1974, page 36.

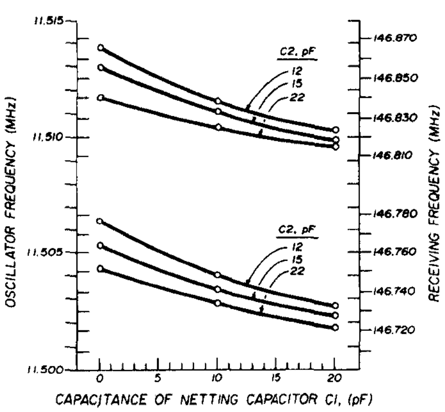


fig. 8. Effect of netting capacitor C1, and capacitor C2 on oscillator and receiving frequencies using a Goral oscillator. Crystals were ground for 11.5050 (146.160) and 11.5125 (146.850) MHz use when operated into a 10 pF load in a GE Progress Line receiver.

simple crystal oven

This unit can be added to existing equipment, such as a frequency counter, that uses a crystal oscillator as a reference frequency. It provides proportional rather than on-off control. All components are mounted on the crystal with several advantages. All heat produced (2 watts maximum) is used in maintaining the crystal temperature, so power consumption is low. All oven components except the trimpot operate at the crystal temperature, aiding stability. Parts layout and schematic are shown in **fig.** 9.

Installation is simple. Note that the TO-5 transistor case is at ground potential. The crystal socket and a simple tie around the foam insulation are sufficient support because of the light weight.

The thermistor, which was from a transistor amplifier bias circuit, is about 1k at room temperature. Values much different from this might require circuit changes. For correct operation the current through the thermistor (about 1 mA) should be much more than the base current of transistor Q1 (0.1 mA). Q1 and Q2 should have low leakage currents. If Q2 is a silicon type, increase the 150-ohm resistor to 680 ohms.

The supply voltages may be available in the existing equipment power supply, and maximum drain is only 200 mA. An unregulated voltage higher than 9 volts would require higher-value heating resistors. Power transistor Q2 supplies some of the heat when the operating temperature is reached and proportional control occurs.

Temperature stability depends on the 5-volt supply and the efficiency of the foam insulation, which can be attached neatly with masking tape. Some heat unavoidably leaks through the crystal socket.

The oven should operate a little above the maximum temperature expected inside the equipment, which

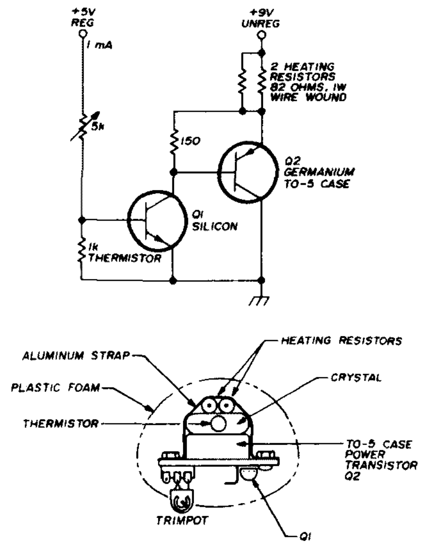


fig. 9. Simple crystal oven.

should be well ventilated. Set the trimpot so the current from the unregulated supply is about 30 mA with the equipment at maximum temperature. My unit reaches operating temperature in 5 to 10 minutes, depending on the ambient temperature.

P. H. Mathieson

notes on 3-400Z, 3-500Z filament circuits

The popular 3-400Z and 3-500Z zero-bias triodes used in cathode-driven amplifiers require 5 volts for the filament supply. A pair will draw 29 amps of filament current. This relatively high current can lead to a series of problems all having the same net result: low filament voltage from excessive resistance somewhere in the filament circuit. Let's assume that in the filament circuit an undesired resistance of only 0.01 ohm is caused by a poor solder connection. At 29 amps, this resistance will result in a voltage drop of 0.29 volt. It will also result in 8.4 watts being dissipated as heat in the already poor connection.

The practical result of this situation was seen in a commercially made amplifier using a pair of 3-400Zs. After about a year of operation, the amplifier gradually started losing output power. The problem has all the appearances of one

or both tubes going bad. The amplifier was upended for a close examination of the filament circuit. The connections from the filament choke had the appearance of cold solder joints. A voltmeter check (performed with the high voltage disabled) showed less than 4.5 volts on the filaments, measured right at the tube pins. A voltage drop of more than 0.1 volt was measured across each of the four solder connections for the filament choke. Enough heat had been generated to soften the solder at each connection. When the four connections were cleaned and resoldered, 5 volts were measured at the tube filaments. Not only was the lost power regained; the amplifier actually put out more power than it had when brand new! In this case the connections were marginal from the start, and enough heat had been generated to continue oxidizing the solder, resulting in a slow deterioration of the connections until finally there was a noticeable degradation in performance.

Every soldered connection in the filament circuit is a potential source of trouble in this regard as are the pressure contacts between tube pins and socket. So if your amplifier has lost some of its pep, check the filament voltage at the tube socket before investing in new tubes. A good solder connection should have no appreciable voltage drop across it, even at 29 amps, but there may well be a drop of 0.1 volt or more across each half of the filament choke, depending on wire size. Prolonged operation with low filament voltage can result in a loss of some of the filament emission. Should this occur, it may be restored by operating the tube at normal filament voltage with no drive or plate voltage for about an hour.

If it appears that the filament has burned out, try resoldering the filament pins and you may be pleasantly surprised. It's not unheard of for the tube pin to develop a poor solder connection and open up completely. Another tube problem that occasionally occurs is a filament-to-grid short.* If this happens while the tube is still in warranty, of course it should be returned to the manufacturer for a replacement. Even if the tube is out of warranty, the situation is not necessarily hopeless. Often it's possible to burn out the short by discharging a large capacitor through it. Measure the resistance from each end of the filament to the grid. Use the filament pin that gives the lower reading, which will allow more current to flow and is more likely to melt the short. One 3-400Z was successfully repaired in this manner four times using a fully charged 2500 μF 200-volt capacitor. Each time this was done, no doubt some of the thin wires in the grid structure were damaged. On the fifth occasion the filament burned out, but the need to purchase a new tube was successfully delayed for over a year.

John E. Becker, K9WEH

*A filament-to-grid short can occur if the filament is allowed to sag against the grid. When upending the chassis to make voltage measurements, make certain the tube is oriented vertically. editor

NEW products

144- and 220-MHz fm transmitters

VHF Engineering has just announced two vhf fm transmitter kits for 144 and 220 MHz, the TX-144B and TX-220B. These new kits offer state-of-the-art design using pre-wound coils, epoxy-glass circuit board, temperature-compensated crystal trimmer, and have a nominal output of 1.5 watts.

Both construction and tune-up have been simplified through the use of pre-wound coils and tune-up test points, making these units one-evening projects. Special tools or equipment are not required. Only a low-wattage soldering iron, solder, wire cutters, and long-nosed pliers are needed for construction. Tune-up requires only a VOM, a small light bulb, and a non-metal hex tuning tool.

The basic transmitters are offered as single channel units, but may be multi-channeled through the use of a simple switch or by using the inexpensive ten-channel option. Companion receiver kits and power amplifier kits are available so that the average amateur may build his own vhf fm transceiver at a very nominal cost. These units may be ordered directly from VHF Engineering, 320 Water Street, P.O. Box 1921, Binghamton, New York 13902 or from one of their many dealers throughout the country. The TX-144B and TX-220B kits sell for $29.95 each plus shipping. New York State residents should add sales tax.

programmable-memory keyer

If you're a contest operator you'll be interested in the MK-75 keyer by Brown and Simpson Engineering of Ontario, Canada. The MK-75 features design innovations especially tailored for serious competition work. The best features of the Accu-Keyer and the TO-type keyer have been incorporated into the MK-75, making it easy to use for those familiar with either of these keying methods.

The MK-75 features self-completing characters, dot/dash memory (Iambic), as well as automatic letter and word spacing. A sidetone oscillator is also provided, with volume and tone controls. Speed range of the MK-75 is 5-65 wpm.

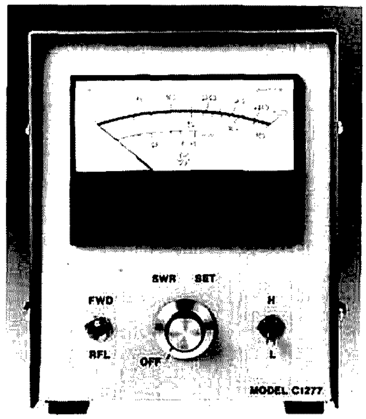
The MK-75 memory is quite versatile. To program a message, you place the READ/WRITE switch in WRITE position, press any one of the four quadrant buttons with its LED display, key in your message, and return the switch to READ position. To read a message, just press a quadrant button.

Suppose you wish to enter an insert into the preprogrammed message. Merely *program* the first part of the message as described above, press the INSERT button, program the remainder of the message, and return the READ/WRITE switch to READ. The memory is now ready for your insert message. To insert a message, such as W8XYZ NR 682, just press the appropriate quadrant button. The memory stops at the point where you wish to insert the message and waits while you insert the message manually. The memory then finishes the preprogrammed message after your insert.

The MK-75 includes many other features such as insert-function bypass, disable or delay of automatic restart capability, instant message interrupt, and message editing.

These are only the basic operations of the MK-75; the user's manual has more detailed information. Price is $249.00, which includes shipping and handling in the U.S.A. and Canada. More information is available from Brown and Simpson Engineering, 17 South Edgeley Avenue, Scarborough, Ontario, Canada M1N 3K9, or use *check-off* on page 110.

rf power and swr meter



The model C1277 broadband power/swr meter by Werlatone is an inexpensive instrument for amateur use in the hf and vhf range. The C1277 covers 27 to 450 MHz continuously and features dual power-range scales for 15 and 50 watts. A unique broadband coupler provides a useful bandwidth approximately eight times greater than previously available. No plug-in units or separate indicators are required. Wattmeter accuracy is ±10% when used with a 50-ohm antenna system. Sensitivity for swr measurements is less than 5 watts. ICAS power capability is 50 watts CW, 27 to 200 MHz; 25 watts CW, 200 to 300 MHz, and 15 watts CW, 300 to 450 MHz. Single-sideband power capability over the entire range is 50 watts maximum.

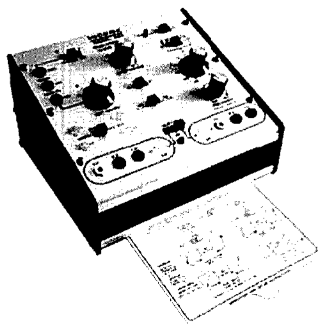
The model C1277 is attractively packaged in a 4x4x5 inch (10x10x12.5cm) enclosure and is equipped with a two-color, wide-view meter. The wideband directional coupler, which is weather-tight, may be removed from the enclosure for remote location.

More Details? CHECK—OFF Page 110

No other environmental protection is needed.

The wattmeter is an in-line instrument and is a useful addition to your station for monitoring transmitter and antenna performance. Amateur net price of the model C1277 is $89.50. If you'd like more information, write to Werlatone, Inc., Brewster, New York 10509, or use *check-off* on page 110.

semiconductor curve tracer



Hickok's new model 440 curve tracer, with exclusive *Insta-Beta* display, dynamically tests all types of semiconductors under actual conditions – in or out of circuit. Used with any scope having an external horizontal input, it generates calibrated characteristic curves that can be accurately scaled right from the screen. It safely tests jfets, mosfets, diodes, zeners, transistors, UJTs and SCRs silicon or germanium, power or signal.

Insta-Beta takes the guesswork out of transistor beta and fet parameter calculations. In the transistor mode, *Insta-Beta* displays a single, full range I_C/I_B curve from which ac and dc beta can be instantly determined without interpolation. This curve also shows beta linearity at a glance. In the fet mode, *Insta-Beta* displays the entire transfer curve including pinch-off voltage, full-on current, and active portion for easy calibration of transconductance.

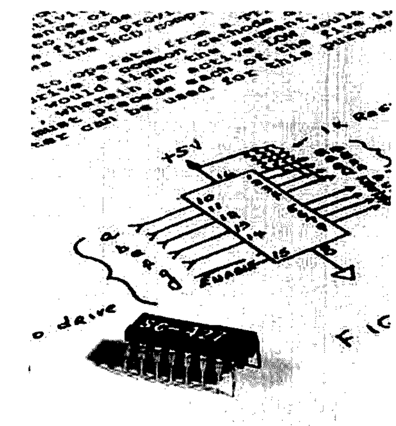
In normal semiconductor testing, a variable step control provides characteristic curve displays with up to ten steps per family (steps of base current for transistors and steps of gate voltage for

fets). Maximum sensitivity of 1 volt per division is especially useful for measurements in the semiconductor threshold or turn-on region.

Controls are logically arranged on the front panel and use color coding and fast set-up marks where applicable. A handy pull-out card provides ready reference information for calibration, set-up and operation of the instrument.

For more information on the Model 440 Curve Tracer contact Tom Hayden, Instrumentation & Controls Division, Hickok Electrical Instrument Company, 10514 Dupont Avenue, Cleveland, Ohio 44108 or use *check-off* on page 110.

decoder ic



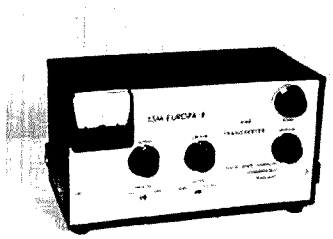
The SC-427 is a new decoder chip available from Scarpa Laboratories in a standard 16-pin DIP which accepts TTL conditioned inputs from a seven-segment display driver and converts them back onto a BCD output.

The device was designed to take advantage of the powerful computing capability of low-cost calculator chips which, in their present form, dead end into a visual digital display. By converting back into BCD format, the engineer is able to break out this extraordinarily economical data-reduction ability into useful computer, controller, time-clock or print-out functions. The unit can also be used to interface LSI clock chips to computers, controllers or printers.

The TTL-Schottky device operates from a single 5 volt supply and has a conversion speed of 25 nanoseconds, thereby requiring only one device for multiplexed displays. For more information, write to Scarpa Laboratories, Inc., 46 Liberty St., Metuchen, New Jersey 08840, or use *check-off* on page 110.

More Details? CHECK—OFF Page 110

vhf transverter

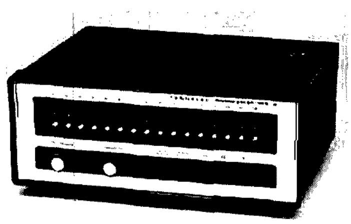


The new Europa B from Solid State Modules is a linear transmit and receive converter from 28-30 MHz to 144-146 MHz or 50-52 MHz and is suitable for use with either a transceiver or separate receiver/transmitter. It is ideal for Oscar operation as well as normal tropo work. A crystal switch and extra crystal can be installed to extend the frequency coverage. Although designed primarily for ssb operation, the Europa B will receive and transmit any mode which the hf equipment is capable, ssb, a-m, fm, FSK or CW.

The receiver converter is broadbanded to cover the entire vhf band without any tuning. It uses dual-gate mosfets for optimum sensitivity, gain and low cross-mod. The noise figure is 2 dB; converter gain is 30 dB. The transmit converter uses tubes to provide high power, good linearity and high rejection of spurious signals. Power input is 200 watts (50% efficiency, minimum), drive requirement, 200 mW. An optional ac power supply is available.

For more information on the new Europa B Vhf Transverter, write to Solid State Modules, 1624 Kaweloka Street, Pearl City, Hawaii 96782, or use *check-off* on page 110.

fm scanning receiver

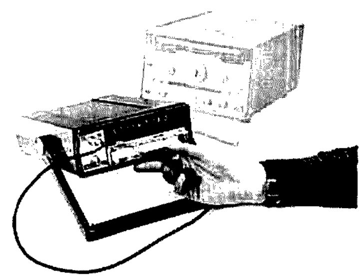


Tennelec, Incorporated, is now offering an improved version of their Memoryscan fm scanning receiver, the Memoryscan MS-2. With this receiver you can monitor up to 16 low/high vhf and uhf channels without buying expen-

sive new crystals. Memoryscan MS-2 has been updated to include new fet rf stages and an effective new filter, all of which mean higher sensitivity, better adjacent-channel isolation, and cleaner, crisper audio.

The MS-2 may be programmed to receive any of 4000 frequencies by punching the appropriate code, listed in Tennelec's code book, into the scanner. Manual scan override is provided, as well as channel lockout switches and squelch. Visual frequency verification capability allows you to check whether the proper code has been programmed. An optional mobile kit is also available. The MS-2 retails for $339.95. More information is available from Tennelec, Incorporated, 601 Turnpike, Oak Ridge, Tennessee 37830, or use *check-off* on page 110.

frequency counter



A portable high-sensitivity frequency counter designed specifically for telecommunications applications was recently introduced by the Fluke Counter Division. This all-new frequency counter, Fluke model 1920A, incorporates many new and innovative features including advanced LSI/MOS circuitry which makes a major contribution to the counter's exceptional electrical specifications while permitting a significant reduction in the unit's size and weight.

The 1920A features a nine-digit LED display, sensitivity to 15 mV, AGC standard, and a frequency range from 5 Hz to 520 MHz. Optional internal prescalers to 1000 MHz and 1250 MHz cover uhf television, 900-MHz telecommunications, and TACAN/DME.

Direct and prescaled inputs are color-coded to match their corresponding function switches to facilitate operation, while the large, seven-segment nine-digit LED display incorporates full leading zero suppression, automatic an-

 More Details? CHECK—OFF Page 110

nunciation, overflow and a self-check mode which lights all digit segments.

Measurement delays have been eliminated in the 1920A through a "rapid-access gate" which free runs in the absence of input signals to be in a position to open the gate for the selected gate time as soon as a signal is sensed. An auto-reset circuit initiates a new measurement every time any front panel switch is activated, ensuring that the first measurement obtained is always correct.

In addition to normal frequency measurements, a burst function switch is provided, permitting the measurement of rf bursts having a duration of 2 ms or more. To avoid erroneous reading, the display is automatically reset to zero if the burst width is less than the gate time selected. An optional resolution multiplier is available which coherently multiplies audio tone signals by 1000, providing a resolution of 0.001 Hz in 1 second.

The 1920A frequency counter is backed by Fluke's full warranty and coast-to-coast service, and is priced at $859, FOB Buffalo, New York. For more information, write to John Fluke Mfg. Co., Ltd., Counter Division, Post Office Box 1094 Station D, Buffalo, New York 14210, or use *check-off* on page 110.

QSL display album

To organize, display and protect your QSL cards, Ace Art Company offers the NuAce QSL Card Display Album. Available in blue, black or ginger binder, the album has 23 chrome steel rings which hold up to 25 pages or 150 cards. Crystal clear pages of durable vinyl give protection from handling, dampness, and fading. Each page has 3 pockets sized 3-7/8 x 7 inches (9.8 x 17.8 cm) and will hold six cards back to back. (Also available is a two-pocket page with 5-5/8 x 7-inch pockets (14.3 x 17.8 cm) for oversize cards, and a one-pocket page sized 12 x 7 inches (30.5 x 17.8 cm).

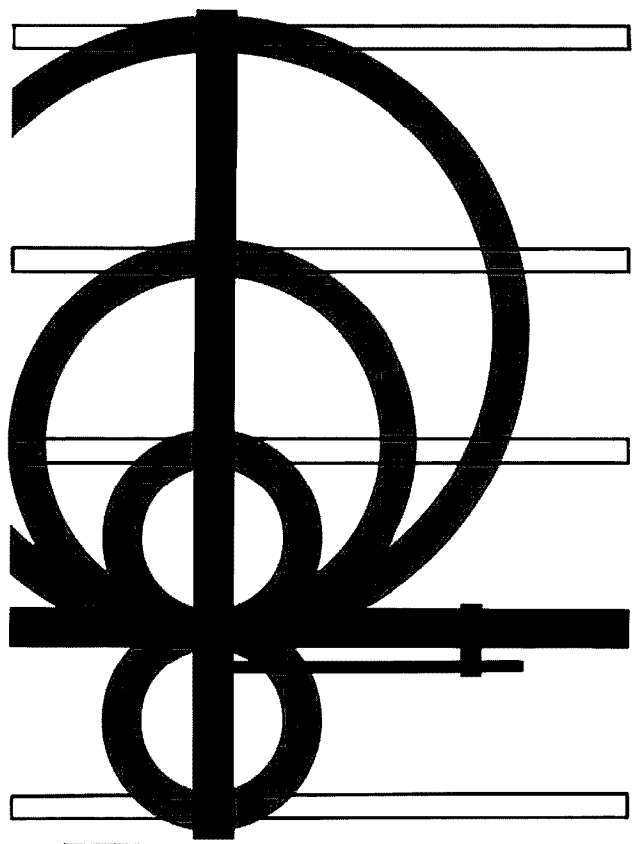
The binder and 10 pages holding 60 QSL cards may be purchased for $5.95 (plus $1.50 shipping and handling) from Ace Art Company, Inc., 24 Gould Street, Reading, Massachusetts 01867. Extra vinyl pages are 49 cents each.

one dollar

ham radio

magazine

MAY 1976



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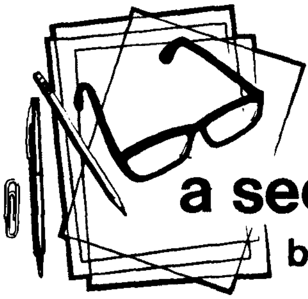


a second look

by Jim Fisk

Although many amateurs still look on the citizens band with a certain amount of scorn, attitudes are changing, and I think it's time that we all took a second look at the CB service and what its tremendous growth in recent months means to amateur radio. There's little doubt that many of the early CB operators were frustrated amateurs, unable (or unwilling) to spend the time *and effort to pass the amateur exam. The* CB ticket provided a painless way to get on the air, to play with radio, to work skip DX, to exchange QSL cards—radio was as much a hobby to them as it is to amateurs, and perhaps it was the CBer's free ride that was so irritating. That, and the fact that they were operating on frequencies which had been expropriated from the amateur service.

This all began to change in early 1974 when the truckers started using citizens band in their much publicized revolt against the 55-mph speed limit. The truckers soon discovered that citizens band offered them a chance to keep track of their pals, to avoid road hazards and traffic snarls, and to relieve the boredom of long hours in the cab of an 18-wheeler. The idea caught on quickly, and it wasn't too long before many travelers started installing CB sets in their cars before they started on a long trip.

The CB service, which had taken sixteen years to grow to a million licensees, quickly doubled, then tripled, as license applications poured into the FCC offices at the rate of 500,000 per month. The new CBer, now the vast majority, wasn't interested in radio as a hobby, but as a medium of communications. The old CBer, crowded out of the band by six-million new users, is still interested in radio as a hobby, and there is growing evidence that many are turning to amateur radio as an outlet. This is a huge potential resource for amateur radio, one that can no longer be ignored.

While many CB operators have been expecting some relief in the form of more channels in the 27-MHz band, the FCC recently announced that CB expansion will be delayed until new technical specifications for CB equipment can be developed. This means that there will be no expansion of the class-D service (or inauguration of the proposed class-E service) until at least early 1977. License applications are still pouring into the FCC at an unprecedented rate, so band crowding is going to get much worse than it is now. There's little doubt that there will *eventually* have to be some sort of relief in the form of expanded CB bands, but that may be more than a year away. This action promises to hasten the immigration of CB hobbyists into the amateur ranks.

As most amateurs know, our bands will come under close scrutiny at the World Administrative Radio Conference (WARC) in Geneva, in 1979. The size of the amateur service has remained practically static for the past ten years or so, and insiders who should know have repeatedly stressed the need for substantial growth between now and 1979. Without growth we're liable to be facing the complete demise of amateur radio as we now know it. The citizens band is a readily available source for that needed growth. The displaced CB hobbyists have already displayed an interest in radio communications and a willingness to equip their stations with top-quality equipment; it's our job to sell them on the idea of becoming radio amateurs. With the present crowded conditions on the CB channels, it shouldn't take much more than a nudge. And when a CBer expresses interest in amateur radio, don't turn your back on him -- he and his friends may hold the key to the whole future of amateur radio.

Jim Fisk, W1DTY
editor-in-chief

presstop

SIGNIFICANT CONFLICTS FOR FREQUENCIES appear in the FCC's first tabulation of the various services WARC Working Group proposals. In a 128-page Public Notice Number 62477 released on March 22, the commission carefully pointed out that at this point no attempt had been made to reconcile conflicts nor were comments on the compilation being sought from the public.

Other Than A 5 kHz Slice off the bottom of the present 160 band, the only challenge to present Amateur Radio allocations below two meters was the Broadcast Service's bid for 3.9-4 MHz. However, all of the proposed new bands and extensions of the present bands are frequencies that other services are also after.

Two Meters Presents a more serious problem, with Aeronautical Mobile looking to pick up 146-148 MHz — a change that would make Region II conform with the two meter allocation in the rest of the world! 220 MHz is being chased by the General Radio Service (CB), but 420-450 MHz was not challenged.

Still Higher In The Spectrum we have several competitors for the 900 MHz segment the Amateur Radio Working Group had proposed, while several of the proposed or present microwave Amateur allocations may also have to be fought for. In view of the rapid upward move of technology, these may become far more important to us in the next decade or so than will be apparent in 1979.

CB EXPANSION DELAYED UNTIL at least early 1977. In a surprise move on March 19th the Commissioners agreed to postpone action on both Class-D (27 MHz) expansion, Docket 20120, and the proposed Class-E (220 MHz) service, Docket 19759, until new technical specifications for CB equipment can be developed. These technical specs will be the subject of a new Notice of Proposed Rule Making which will propose tight limits on both receiver incidental radiation and transmitter spurious and harmonic outputs. The Commissioners set the end of this year as the deadline for action on this as yet unwritten NPRM, and said that until it is acted on no further action will be taken on either CB expansion docket. At the same time they also said they plan to include the recently announced study to be conducted by FCC's Office of Plans and Policy on the "Long Term Needs of the General Public for Personal Radio Communications" in their future deliberations on possible CB expansion.

The Delay Certainly came as a shock to the CB industry, whose insiders had predicted the Class-D expansion to 40 or 50 channels would be announced at the FCC Forum at PC '76 in late March.

"POINT OF MAIL" CB LICENSING was also approved by the FCC and is supposed to have been implemented by late April. Under the new plan the buyer of a CB radio will receive a license application form from either the dealer or packed with the radio. He fills out the form, which he then takes to his Post Office where it will be receipted and mailed for a nominal charge. He can then go on the air immediately with his self-generated call using the Post Office receipt as his Special Temporary Authority to operate.

WRITTEN CW REQUIREMENT for Amateur Radio license examination has been dropped from Part 97 of the Rules in an FCC order released March 12th. The change, which deletes need for "at least 1 minute" of error-free copy became effective March 24, and opens the way for substitution of questions on content as an alternative method of determining CW competence. It won't really change things for most Amateurs for some time to come, however — a few selected FCC Offices will probably start administering "no write" exams on an experimental basis almost immediately, but the rest will continue giving exams with a written code test until a final procedure worked around the new approach can be developed.

RFI ELIMINATION EFFORT IN CONGRESS has received a valuable boost with Senator Barry Goldwater's introduction of his Senate version of Representative Vanik's bill (HR7052). Goldwater's Senate bill, S3033, deserves our strongest support — write your Senators today asking them to help push it and send a copy of your letter to K7UGA at the Senate Office Building.

AMATEUR FAST-SCAN TV REPEATERS will be permitted to operate in the 420-450 MHz Amateur band under a waiver authorized by the Chief of the FCC's Safety and Special Services Bureau. The waiver suspends Section 97.61 (c) of the Commission's Rules, which limits repeaters to 442-450 MHz, for a period of one year — to February 27, 1977 — to permit any licensed Amateur repeater station to conduct fast-scan TV experimentation on the band without having to request prior FCC approval.

HIRAN HAS BEEN APPROVED for operation in the entire 420-450 MHz Amateur band in a Report and Order that became effective April 22. HIRAN, the high accuracy radio location system that was developed for the off-shore oil drilling industry, will use the 70-cm band on a secondary, non-interfering basis to the Amateur service. We also share the band with government Direction Finders.

PIN diode transmit/receive switch for 80-10 meters

Need a fast, dependable break-in system? This design handles up to 100 watts output with keying speeds as high as desired

Popular break-in **methods** include the vacuum-relay and saturated-amplifier systems. The vacuum relay, actuated by a keyer circuit, switches the antenna between receiver and transmitter in accordance with keying commands. This system is effective but is limited to keying speeds of 55-60 wpm. The saturated-amplifier system consists of *an rf amplifier* loosely coupled to the antenna. The amplifier has enough gain to overcome the loss caused by loose coupling to the antenna. When the transmitter is keyed rf saturates the amplifier, decreasing its gain, which provides receiver-transmitter isolation from the antenna during key-down conditions. This system has a number of shortcomings, which are discussed later.

A solid-state TR switch that operates at very high speeds and provides the advantages of the vacuum relay keying system would be a worthwhile addition to the station of the amateur who enjoys working DX and high-speed CW. Such a switch would also be satisfactory for phone work.

design objectives

My objective was to design a low-power, solid-state switch that would transfer the antenna between receiver and transmitter in accordance with transmitter keying commands at the highest speed desired. This TR switch would overcome saturating-amplifier TR switch limitations, which include:

1. Receiving system sensitivity degradation caused by antenna-to-receiver coupling loss.

2. Receiver desensitization caused by noise in the transmitter power amplifier stages.

3. Receiver sensitivity limitations between dots and dashes caused by TR switch saturation during key-down conditions (amplifier recovery time).

4. Sensitivity, isolation, and transmitter loading variations with frequency caused by the mismatch introduced by paralleling the TR switch with transmitter output.

5. Harmonic generation and TVI in extreme cases.

An additional objective was to obtain a TR switching system with losses and isolation approaching a vacuum-relay system but with higher speed capability.

PIN diode basics

The PIN diode is the heart of this TR switch. Its key parameters are illustrated in **figs. 1 to 3**, which show the impedance characteristics of a typical Unitrode UM4000-series device versus forward-bias current and reverse-bias voltage. With a forward dc bias current of about 35 mA, (**figs. 1** and **3**) the diode looks like a 1-ohm resistor below 500 MHz. With a reverse bias voltage of, say, 100 volts dc, (**fig. 2**) the diode looks like a 2.5 pF capacitor in parallel with a 60k resistor at 100 MHz and an even higher resistance at lower frequencies.

These characteristics mean that a single PIN diode can be made to provide 0.1 to 0.2 dB insertion loss when forward biased and about 25 to 30 dB isolation when reversed biased. A PIN diode also looks like a relatively high impedance without reverse bias and no forward direct current.

PIN diodes have sometimes been considered microwave devices, unusable at hf. However, the lowest frequency at which a PIN diode exhibits the characteristics of **figs. 1** to **3** is related to "carrier lifetime," which is usually expressed in fractions of a second. Long-carrier-lifetime diodes will exhibit these impedance/bias characteristics at lower frequencies than short-carrier-lifetime (microwave) devices, which behave like regular diodes at lower frequencies. The PIN diodes illustrated have typical carrier lifetimes of 7 microseconds, which makes them useful in this application below 1 MHz to over 3 GHz.

In a TR switch application the diode peak inverse voltage (PIV) rating limits the reverse bias that can be applied to isolate a source from a load, and the maximum junction temperature limits the amount of rf power that can be connected between source and load.

By James K. Boomer, W9KHC, 4031 Dalewood Drive, Fort Wayne, Indiana 46805

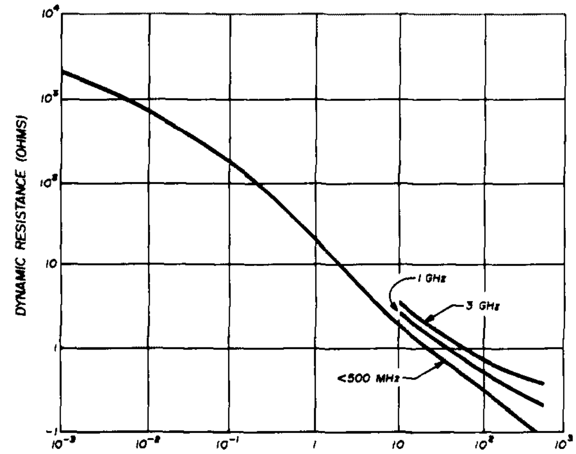


fig. 1. Typical forward resistance characteristic, UM4000-series PIN diode.

The Unitrode UM4000-series diodes have PIV ratings to 1 kV (UM4310) and maximum junction temperature ratings of 347°F (175°C).

functional description

During key down CR1 (fig. 4) is forward biased, connecting the transmitter to the antenna. CR2, CR4 are reverse biased and CR3 is forward biased, isolating the receiver from antenna and transmitter. For best operation CR2 reverse dc bias should be at least as large as transmitter peak rf output voltage, which prevents distortion or rectification in CR2 during key-down operation.

During key up CR1 is zero biased (or reverse biased if desired) to isolate the transmitter tank circuit and noise generated in the transmitter from the antenna and receiver, thereby preventing receiver desensitization. The receiving-system impedance match is also preserved. CR2, CR4 are forward biased and CR3 is reverse biased, thereby connecting the receiver to the antenna.

The three diodes in the receive path provide a theoretical isolation of 75 to 90 dB between receiver and antenna terminal during transmit. This isolation corresponds to about 6 mV applied to the receiver with a 50-watt output transmitter, assuming a 50-ohm system, or about 8.5 mV open-circuit voltage at 100-watts transmitter output. These key-down receive signals are well within the dynamic range of most amateur receivers. Fewer or more diodes may be used in the receive leg, depending on desired receiver isolation.

circuit description

Fig. 5 is a schematic of the TR switch, which is designed for power levels to 100 watts. Normal (nominal) dc voltages are listed on the diagram. CR1 is forward biased for about 45 mA dc in transmit and CR2 is reverse biased by 124 volts in transmit. These two conditions ensure minimum loss in CR1 and maximum isolation in CR2 during transmit.

CR3, CR4, shown as UM4004s, could probably be replaced by less-expensive 1N5767s or possibly even high-grade computer switching (high-conductance, low-capacitance) diodes. The 1N5767s, which are also PIN diodes, will give equivalent isolation whereas the computer diodes may not, because of their possible larger capacitance.

Fig. 5 is designed to operate from the collector of the Q3 keying-circuit transistor of the Touchcoder II,[1] which is shown in dotted lines. However, any source providing the T (transmit) and R (receive) voltages shown will key the circuit.

transmit operation

The voltage to R4, Q1's base resistor, is low (0.2 volt), causing Q1 to be off, which results in a high voltage at Q1's collector. This, in turn, turns on Q3, Q2, allowing CR1 to conduct about 45 mA of forward current, permitting transmitter power to flow to the antenna. In additon it causes CR3 to conduct, thereby presenting a low rf impedance (CR3 conducting in conjunction with bypass capacitor C9). The current through CR3 causes sufficient voltage drop across R2 to reverse

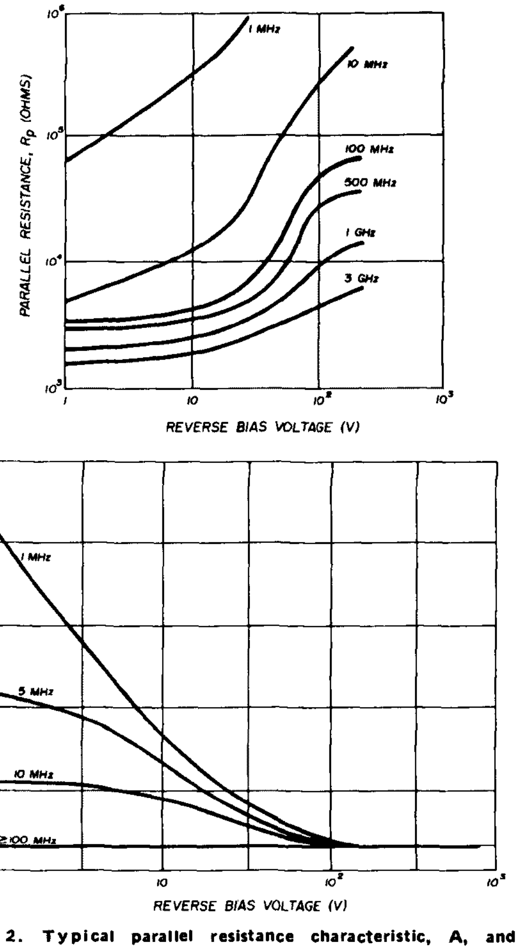


fig. 2. Typical parallel resistance characteristic, A, and capacitance characteristic, B, UM4000-series PIN diode.

bias CR4 (voltage at CR3, CR4 junction is less than the fixed bias provided by CR5, CR6). This, in turn, causes CR4 to present a high rf impedance.

receive operation

The voltage to Q1's base resistor, R4, is high (2 volts) causing Q1 to be on (saturated), which results in a low voltage at Q1's collector. This, in turn, causes Q2 to be

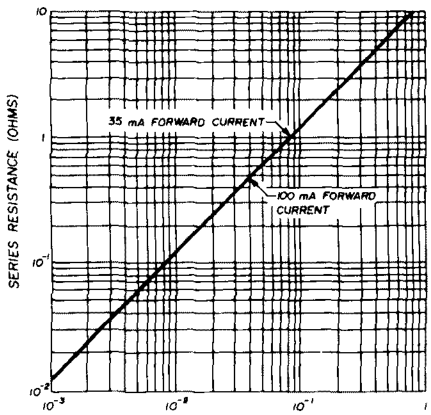


fig. 3. Insertion loss versus series resistance (one diode in series with 50-ohm load).

off, resulting in a high voltage at Q2's collector. This condition allows CR2 to conduct through R1, Q1, thereby presenting a low rf impedance.

The low voltage at Q1's collector also causes Q3 to be off, *which renders* CR1 essentially zero biased and CR3 reverse biased, resulting in high rf impedances in the two diodes. The absence of current in CR3 removes this com-

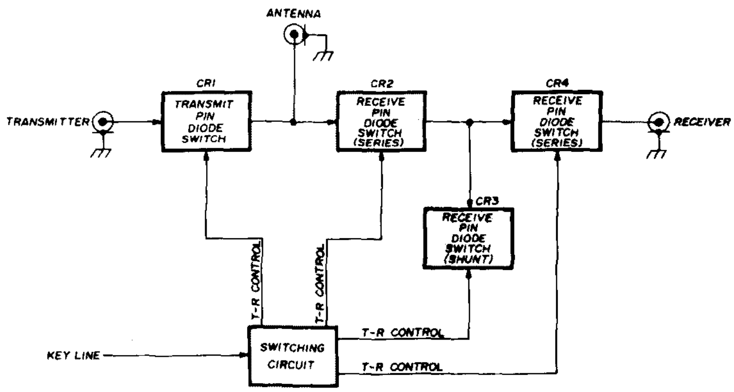


fig. 4. Simplified functional diagram.

ponent of voltage from R2, permitting current to flow through R2, CR4 and CR5, CR6. Thus CR4 looks like a low rf impedance. *Diode* CR7 *ensures that* Q3 *will* be off when Q1 is on.

component selection

The rf chokes must have no series resonances in the bands of interest. Standard pi-wound 2.5-mH units usually look like about 50k-100k ohms resistance in parallel with 1-2 pF between 3.5 and 28 MHz. In other words, all rf chokes must look like high impedances in the frequency range of interest otherwise they will absorb rf power and cause losses in receive and possibly burn up in transmit. It's well to measure the rf impedance of these chokes on an rf impedance meter. Also, L1, L2 should be rated for at least 50 mA for the circuit in **fig. 5**.

Capacitors C1-C9 should be good-quality, low-inductance ceramics. Note that C1 and C2 carry the transmitter rf output current. Transistor Q1 must have sufficient voltage rating to switch up to 124 volts dc. Q2 need not be a high-voltage unit (I used 2N5550s, which I had in the junk box). Q2 could be a low-voltage switching transistor *(high beta and low* leakage are preferred to ensure good switching characteristics).

The power supply shown in **fig. 5** has minimum acceptable filtering (the junk box was almost empty); thus additional filtering may be desired. Listen for ripple on received signals and add capacitance to suit yourself.

The TR line that switches Q1 could also be *used to* activate additional logic to reduce receiver rf gain if a better monitoring note is desired. However, the unit shown gives a reasonably good monitoring note with my DX-60A transmitter output level.

keying time constants

The circuit shown works well with my transmitter, which has a hard keying characteristic; however, keying time constants are mentioned briefly for completeness.

When the key is depressed the circuit will switch to transmit instantaneously, and in any reasonably shaped transmitter keying system, the circuit should be in the transmit mode before the transmitter rf output appears at J1. When the key is released, the circuit will also return to receive instantaneously, because no "deliberate" time constants are built in except for the inherent time required for Q1 collector to change from 124 to 0.8 volts dc.

A transmit-to-receive delay can be easily obtained by adding capacitance on the TR input line (line feeding

R4). This added capacitance will have virtually no effect on receive-to-transmit transition time, because the keying transistor (shown as Q3 in fig. 5) will instantaneously discharge the capacitance when driven into conduction with key down. However, when the key is released the keying transistor base current will cease instantaneously,

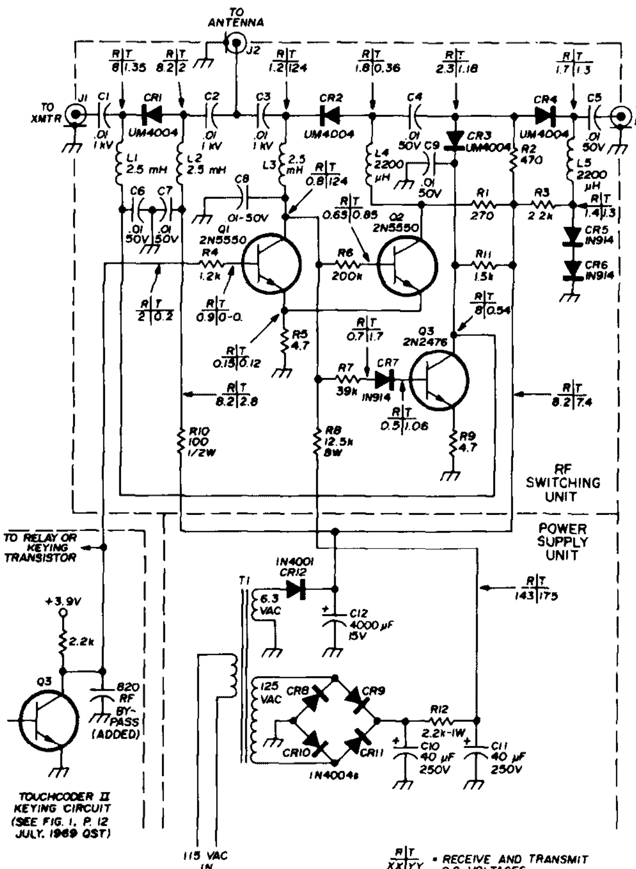


fig. 5. PIN diode TR switch schematic.

but the collector voltage rise to receive level will be delayed by the time constant of the 2.2k resistor in Q3's collector and the added capacitance on the TR switch input line.

construction

The photos show the completed 100-watt TR switch, which is mounted in a 3- by 4- by 5-inch (7.6- by 10.2- by 12.7-cm) Minibox. What looks like extra parts, compared with the schematic, are components that were paralleled to obtain desired values. For example, the power supply contains some paralleled capacitors to obtain the power-supply output capacitance, and two resistors were paralleled in two cases in the rf switching unit to obtain desired values.

A full-size drawing of the rf switching unit PC board is shown in fig. 6. This figure is included primarily to illustrate the general layout of CR1-CR4, L1-L5, and C1-C9. Note that these components are laid out essentially as shown in the schematic. Chokes L1-L5 should be oriented at right angles to each other to minimize mutual stray coupling (although I didn't do this). Note the ground plane strip extending the length of the board in the drawing, which provides a low-impedance grounding strap for C6-C9. The layout shown is only a guide — it was made to fit the parts on hand and can be modified to suit your preference, providing the guidelines mentioned above are followed. I used a low-cost Vector board kit as a source of board and fabrication materials.

All capacitor and PIN diode leads should be as short as possible to ensure proper rf impedance characteristics. Take care not to overheat the diodes and other semiconductors when soldering them to the board. The ARRL handbook is a good reference for PC-board construction.

final assembly

Ground lugs are placed under the rf connector retaining nuts (on the inside of the box), and are positioned and bent to permit soldering the rf switching unit PC board ground plane edge strips to them, thereby fastening the rf switching unit into the box. The rf connector center pins are then soldered directly to the pads on the rf switching unit board. These center pins can also be connected to the board by short lengths of large bus wire. AWG 18 or 16 (1.0 or 1.3mm) may be used, depending on the rf connectors chosen.

The power supply can be mounted as shown or mounted underneath the rf switching unit on the main part of the Minibox. Power and TR switching interconnections could be made through a connector and cable (cable mounted at the front of the Minibox). In this case, the rear cover would not contain the power supply but would be just a simple cover.

checkout and operation

Inspect the unit for wiring errors and solder splashes or other solder bridges between PC-board pads, and check all solder joints for quality. Apply ac power and check all voltages against the values shown in the schematic (receive and transmit). These are nominal values and may vary slightly. The important voltages are the voltage drops through CR1-CR4 rather than absolute values at each diode terminal. The cover can now be attached and the unit connected to the transmitter, receiver and antenna.

I used my Touchcoder II keying circuit to activate the TR switch. Other acceptable methods for applying transmit and receive control signals for Q1 include a) a set of contacts on a keying relay that will supply the required voltages, and b) a transistor-level converter from an existing keying circuit. In operation, the TR switch will switch the antenna between transmitter and receiver in accordance with keying commands. The receiver is connected to the antenna between dots, dashes, and words allowing full receiver sensitivity during these periods (limited only by receiver recovery time).

Keying time constants can be checked by using a dual-trace scope, with one trace monitoring transmitter rf output and the other individually monitoring voltages

at the junctions of L2, C7 and L3, C8 (and elsewhere if desired).

TR switch performance was measured on 80 through 10 meters with the following general results:

1. Loss between transmitter and antenna in transmit was about 0.2 dB.

4. Isolation between antenna and receiver in transmit was greater then 78 dB on 80 meters to 71.5 dB on 10 meters.

I have made a preliminary analysis and developed a concept for a high-power, 1-kW output solid state TR switch; however, no circuits have been built to date.

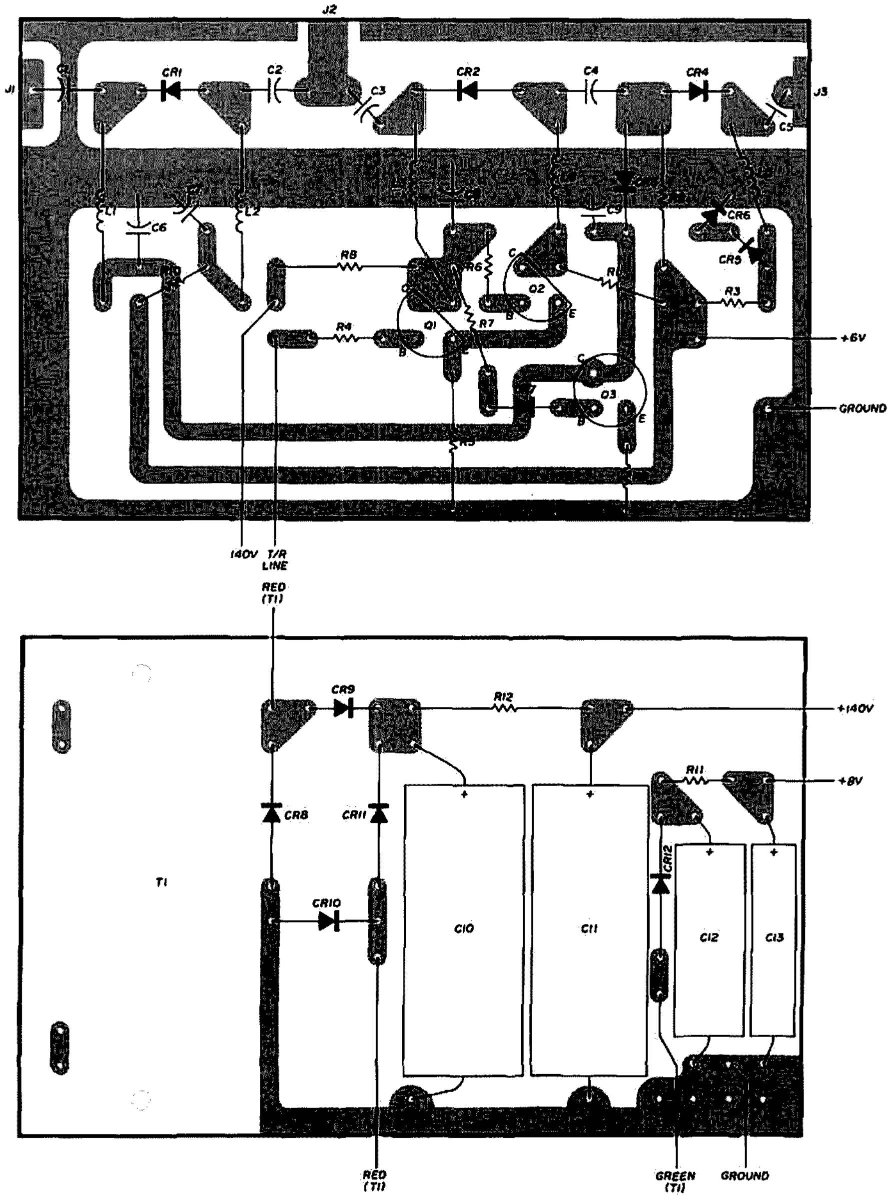


fig. 6. Suggested layout for the rf switching unit, above, and power supply, below.

2. Loss between antenna and receiver in the receive mode was 0.8 dB.

3. Isolation between transmitter and receiver in receive mode was 22 to 12.5 dB. The lower isolation occurred on 10 meters, as expected, but no transmitter noise or loading were noted on 10 meters.

Some critical components and ratings that must be considered in such a TR switch (see fig. 5) are the voltage and current ratings of C1, C2; voltage rating of C3; power rating of CR1; PIV rating of CR2; voltage rating of Q1; and quality and ratings of L1-L3.

For high power, it's almost imperative that a design be used that doesn't require capacitors C1 and C2, be-

cause low-reactance, high-current, high voltage capacitors are very scarce, if available at all. At 1-kW output the rms rf current is about 4.5 amperes at a 50-ohm impedance level. For a 1-kW output TR switch, C4 must be a good-quality ceramic capacitor with at least a 1-kV rating. PIN diode CR1 must be capable of passing the 4.5 amperes of rf current with the lowest possible loss.

Referring to figs. 2 and 3, it is seen that CR1 will look like an rf impedance of about 0.5 ohm with a forward current of 100 mA below 500 MHz, which will result in a CR1 insertion loss of about 0.04 dB in a 50-ohm system. This means that the diode will dissipate about 10 watts at 1-kW output level.

Unitrode makes a stud-mounted series of diodes (UM4000D series) that will dissipate 20 watts at room temperature, which provides adequate margin. The devices will also handle 4.5 amperes of rf (the UM4001D or UM4004D are probably good choices). The stud on these diodes is insulated from the signal leads; thus they can be directly mounted to a good heat sink to provide safe thermal performance.

As noted earlier, best performance is obtained if CR2 is reverse dc biased by an amount at least equal to the transmitter peak rf output voltage. This corresponds to about 316 volts dc in a 1-kW output, 50-ohm system. In systems with high transmission-line swr, the peak rf voltage and current at the TR switch antenna terminal can be higher or lower than its 50-ohm value, depending on swr, line length and operating frequency. For example, if a 2:1 swr exists on the transmission line, the peak rf voltage at the antenna terminal could be as high as 450 volts instead of the previously noted 317 volts. Thus CR2 should have a PIV of at least 350 volts; preferably greater than 500 volts if it is to be reverse biased up to 450 volts during transmit.

Switching transistor Q1 must have voltage ratings compatible with the reverse bias to be applied to CR2 during transmit. Rf chokes L1-L3 should exhibit a high rf impedance over the frequency range of interest (preferably greater than 100k in parallel with 1-2 pF). If these chokes exhibit a low rf impedance at any operating frequency, a high current will flow in them causing possible burnout. L1 and L2 must also have dc ratings of at least 125 mA for 1-kW operation (CR1 forward current assumed to be 100 mA).

transceiver application

This TR switch could be incorporated into transceivers, depending on their design. The design should afford protection of the receiver front end during transmit. The major consideration would be the method of B+ switching incorporated in the transceiver and the degree of commonality of transmitter and receiver circuits; i.e., the degree to which receiver amplifiers are used in transmit and vice versa.

I would like to thank C. H. Glenn for his assistance in selecting PIN diodes for this project.

reference

1. L.Bryant, W4UX, "Touchcoder II," *QST*, July, 1969, page 11.

ham radio

tips on soldering tips

Until a better method than soldering comes along, we're stuck with soldering guns. Here's how to make a wire-bending jig that can be used to make professional-looking, soldering-gun tips that will fit Weller models 8100B, 8200N, D440, and 8100.

You'll need the following items: one new soldering-gun tip; a block of wood approximately 3.5 inches wide by 4.5 inches long by 1 inch thick (9 by 11 by 2.5cm); three 2-inch-long (5cm) finishing nails; five 1-inch-long (2.5cm) brads; and six pieces of no. 12 AWG (2mm) bare copper wire, each piece cut to just 5 inches (12.7cm) long (straight; no bends).

Step 1: Place the new tip on the center of the block with the two tip ends level of flush with the right side of the block. Trace the complete outline of the soldering-gun tip onto the block (see fig. 1). Set the new tip aside.

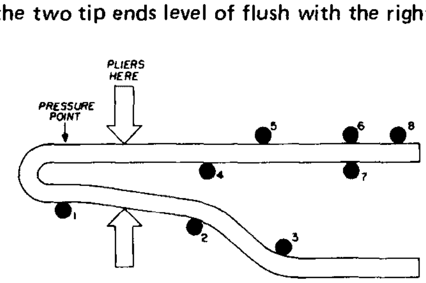


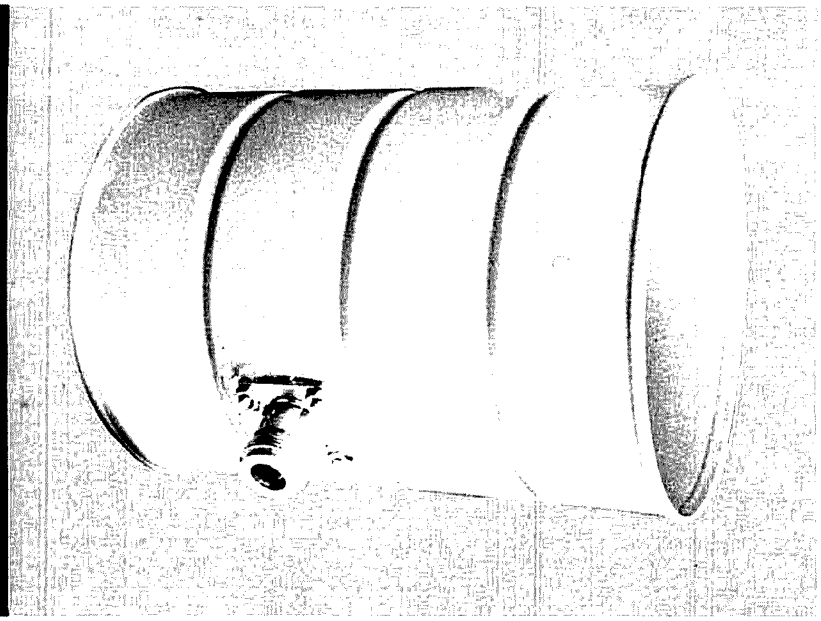
fig. 1. Jig layout for making soldering tips for Weller soldering guns.

Step 2: With a 3/32-inch (2.4mm) drill, bore three holes each 3/4 inch (20mm) deep at points 1, 2, and 3 (fig. 1). These holes will be for the finishing nails to provide bending points on the jig. (A nail is inserted into each hole individually each time a bend is made, then all three nails are removed when the new tip is completed.) Next, drive a brad into the block at points 4, 5, 6, 7, and 8 so that ½ inch (13mm) is straight above the block surface. These brads form a channel for the straight portion of the tip.

Step 3: Lay a piece of the copper wire into the channel with the end of the wire flush with the right side of the jig. Insert one of the finishing nails into hole 1. Hold finger at the pressure point and firmly bend the wire downward around the nail at hole 1 to form a U-shape. Use pliers to make the U tight, keeping the legs of the U parallel but not touching. Next insert the second finishing nail into hole 2 and bend the wire downward toward the bottom right corner. Insert the third finishing nail into hole 3 and bend the wire upward until parallel with the length of wire in the channel.

You now have a soldering-gun tip, however imperfect. Use the remaining wire to improve your skill.

Howard J. Stark, WA4MTH



cylindrical feed horn for parabolic reflectors

Design data with complete construction and tune-up instructions take the guesswork out of building your own feed system

This article provides a simple step-by-step procedure for the design and construction of a feed horn that will work with any parabolic reflector. The objective is to optimize the overall performance of the parabolic antenna system using readily available materials and simple test equipment.

The horn feed has for many years been considered the standard means of illuminating parabolic reflectors.[1] While the rectangular horn has been most often used, more recently circular horns have been used for amateur work.[2] This article explains why the circular horn is a good choice in terms of performance; besides, cans of appropriate sizes and shapes are readily available.

Some pitfalls await the experimenter not familiar with waveguide theory. First, there's an optimum location for the probe that excites the horn. Second, a distinct cutoff frequency is related to horn diameter below which performance rapidly deteriorates. Finally, the choice of feed-horn diameter is important in terms of the focal length/diameter (F/d) ratio of the parabola, because the feed horn radiation pattern depends on feed-horn diameter. It's important that the horn illuminate the parabola effectively.

general considerations

Sometimes relationships that appear to be relatively simple are in fact quite complex. This is certainly the case for the horn antenna. The horns discussed here are relatively short lengths of cylindrical waveguides, shorted on one end. For one thing, what goes on inside a waveguide is unreal compared with the unguided or free-space situation. To put it in Maxwellian terms, the boundary conditions are entirely different. While most amateurs have developed a fairly good mental picture of electromagnetic waves propagating in free space, guided waves are a much more complex matter.

Instead of propagating in straight lines, rf energy moves through a waveguide by bouncing off the walls in a zig-zag manner. Because of interference set up in the guide due to these multiple reflections, the phase of the wave appears to travel faster than the speed of light. As a result, the wavelength in the guide is greater than in free

By Norman J. Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

space.[3] This startling revelation simply means that the wavelength inside the guide is a stretched-out version of the free-space wavelength. The reasons for the stretchout have to do with the so-called phase and group velocities within the guide. For those who wish to pursue this matter further, an excellent and easy-to-comprehend account is given in reference 3.

probe location

If you wanted to transmit two signals of exactly the same frequency down the same waveguide, you would adjust their phase so that the signals would reinforce each other. In a horn antenna, the wave reflected off the closed end of the horn should be in phase with the direct wave traveling out the open end of the horn. The way to accomplish this is to locate the probe one quarter of a guide wavelength from the short. (Note the terminology "guide wavelength.") Under these conditions, it takes exactly one half an rf period for the reflected wave to return to the probe. In the meantime, the driving voltage at the probe has reversed its polarity, but so has the reflected wave in the reflection process. Therefore, the direct and reflected waves are in phase and will reinforce each other. Any other arrangement will provide results that are less than desirable.*

Fig. 1 shows the probe location in a 1296-MHz circular waveguide horn with respect to the horn diameter. The probe location varies with horn diameter because phase velocity depends on the diameter. The graph shows the quarter-wave spacing of the probe from the shorted end of the guide.

Unpainted feed horn made from a 1-pound (.45kg) coffee can

The equation for the guide wavelength is included on the graph of probe location for reference. The curve is based on this equation, using $\lambda_c = 3.42r$, where r is the horn radius and λ_c is the horn cutoff wavelength.

Note that in fig. 1, the smaller the horn diameter, the further the probe should be from the shorted end of the guide. The probe spacing increases rapidly for horn diameters less than about 6 inches (15cm) because the waveguide cutoff frequency is being approached. For

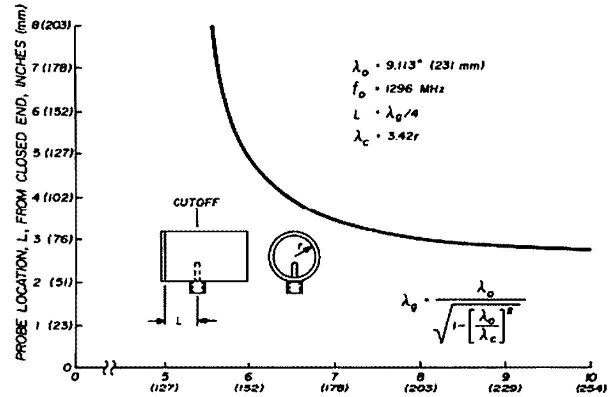


fig. 1. Probe location for a 1296-MHz circular waveguide horn with respect to horn diameter. The equation for the curve is also shown, where λ_o and f_o are respectively wavelength and operating frequency; L is probe location from the closed end in inches (mm); λ_c is horn cutoff wavelength; and r is horn radius.

1296 MHz the cutoff horn diameter is 5.37 inches (13.6cm). At cutoff, rf power does not propagate in the guide. For example, a 2-pound (0.9kg) coffee can with a diameter of 5 inches (12.7cm) will give very unsatisfactory performance at 1296 MHz.

Fig. 1 can be used for 432 MHz simply by multiplying the horn diameter and probe location numbers by 3. Fig. 2 shows probe locations for 2304-MHz horns. The curves can be scaled for use at other frequencies by multiplying the diameter and probe locations by the ratio of the two frequencies. For example, a 6-inch-diameter (15.2cm) 1296-MHz horn scales to 3.375 inches (86mm) at 2304 MHz by multiplying the diameter by 1296/2304. Using the same multiplication factor, the probe (which is 4.9 inches [124mm] from the short on 1296 MHz) should be 2.76 inches (70mm) from the short at 2304 MHz. The overall length of the horn antenna can be any reasonable value so long as the probe is not located immediately at the open end of the horn. A good rule of thumb is to make the horn length between $2L$ and $3L$.

The radiation from an open-ended waveguide has a pattern whose beamwidth varies with the waveguide diameter. It's important to select a horn-feed diameter whose radiation pattern will illuminate the parabola

*Other conditions will provide an in-phase reflected wave; namely, when the probe is located $\frac{n}{4}\lambda$ from the short, where n is any odd integer such as 1, 3, 5 . . . etc. Only the case of $n = 1$ is considered here.

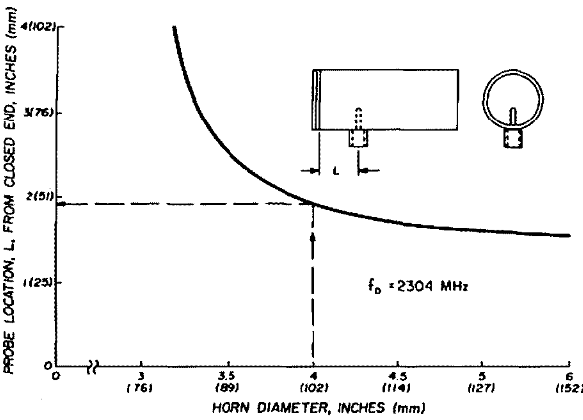


fig. 2. Probe locations for 2304-MHz horns. Curves can be scaled for use at other frequencies by multiplying horn diameter and probe locations by the ratio of the two frequencies.

most effectively. Because of the geometry involved, a parabola with a small *F/d* ratio should be illuminated with a feed horn of small diameter. To illustrate the relationship, a polar diagram of the radiation pattern of a typical horn is shown in **fig. 3**. The relative amount of power being radiated in any given direction is proportional to the length of the radius vector from the focus to a point on the curve.

A cross section of a typical parabola has been superimposed on the polar diagram. Note that less power is directed toward the rim of the dish than toward the center. Furthermore, some of the energy misses the dish entirely, resulting in spillover. If a feed horn having a very narrow radiation pattern is used to reduce spillover, most of the energy will be directed toward the center of the dish with the result that the outer part of the parabola is not used effectively. A condition somewhere between these two extremes represents good design. According to reference 1, overall efficiency peaks out when the illumination at the reflector edge is about 10

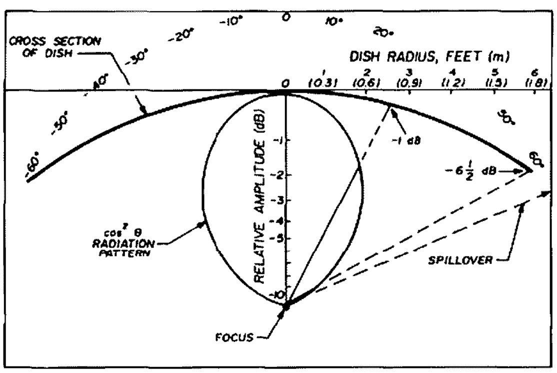


fig. 3. Cross section of a parabolic reflector superimposed on a typical feed-horn radiation pattern. This graph illustrates the importance of selecting a horn-feed diameter whose radiation pattern will illuminate the dish most effectively. Focal length/diameter ratio, F/d, is 0.417; horn diameter is 6.3 inches (16cm), and illumination taper is 6.5 dB. Note that the illumination taper is not ideal.

to 12 dB down from that at the center. Note that the illumination taper in **fig. 3** is not ideal.

The manner in which the beamwidth of a 1296-MHz circular horn antenna varies with horn diameter is illustrated in **fig. 4**. Since the beamwidth of the horn depends on its diameter, it's possible to select a horn diameter to match the particular parabolic reflector to be illuminated.

Fig. 5 is a curve from which the feed-horn diameter can be selected to match the parabola *F/d* ratio. For example, if your dish has a 6-foot (1.8m) focal length and a 16-foot (4.9m) diameter, the *F/d* ratio is 6/16 = 0.375, and the proper horn diameter from **fig. 5** is 7 inches (17.8cm). **Fig. 5** is based on a dish illumination

Plastic coffee-can lid makes an effective rf-transparent radome that's bird and weatherproof.

with a 10-dB taper. According to **fig. 1**, the probe for this horn should be located 3.6 inches (9.1cm) from the shorted end. A reasonable overall horn length would be 7.5 inches (19.1cm).

Fig. 4 is based on average values of vertical and horizontal beamwidths. In practice, assuming horizontal polarization, the illumination will taper more rapidly at the top and bottom of the parabolic reflector than at the sides, because the horn vertical beamwidth is less than the horizontal beamwidth. This situation might be rectified by using a sectoral or elliptical horn. In the latter

case, a circular horn such as described here might be slightly flattened at the open end so that its diameter is increased in the horizontal direction. This will tend to produce a radiation pattern more nearly circular.

feed-horn probe

One of the pertinent factors involved in feed-horn design is the probe for exciting the waveguide. The photographs show a 2304-MHz probe designed for use with a 4-inch (10.2cm) diameter horn made from a 1-pound (0.45kg) coffee can. A 15/16 inch (24mm) length of 0.157-inch (4mm) diameter brass tubing is soldered to the lug of a UG-58A/U type N connector. A 3/4-inch (19mm) length of 0.185-inch (5mm) diameter brass tubing slips over the probe for length adjustment. A 3/8-inch (9.5mm) diameter, 1/4-inch (6.5mm) long brass washer slides over the adjustable sleeve to tune out reactance introduced at the connector.

The connector is attached to the horn at the probe location with brass screws. After the connector holes have been drilled, and before mounting the probe assembly, the paint around the connector flange area should be cleaned off and the flange area tinned with solder.

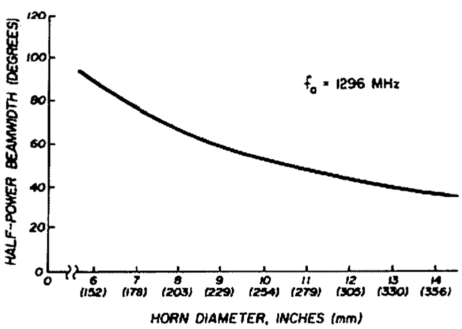
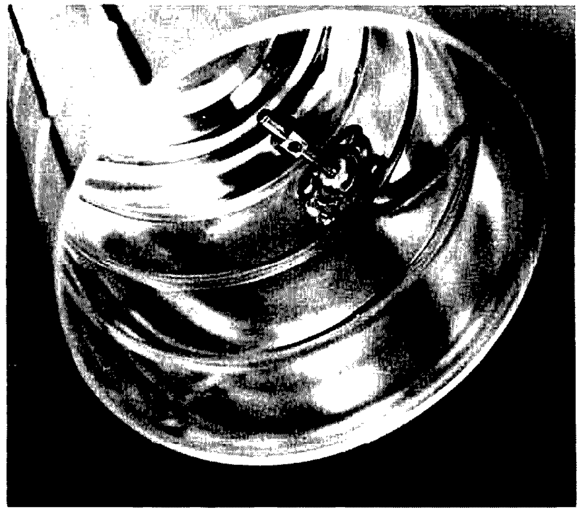


fig. 4. Relationship of beamwidth and 1296-MHz feed-horn diameter. Average values of vertical and horizontal beamwidths are shown.

Once the feed horn has been made, it's only necessary to adjust the probe to put it in operating condition. This can best be done with the aid of a directional coupler and a detector, together with the other components illustrated in fig. 6. Use a directional coupler with a coupling value consistent with transmitter power level. For example, a 60-dB coupler should be used with a 300-watt transmitter; a 20-dB coupler will do the job if the transmitter power is 0.1 watt. In any case, it's desirable that power to the crystal detector be less than about ten milliwatts. The 3- or 6-dB rf pad should be used to ensure a good power match at the coupled arm output.

Orient the feed horn so that energy is not reflected back into the horn while the probe adjustments are being made. Caution: Do not look into the horn when the transmitter is energized, otherwise eye damage can result, particularly if high power is being used. It is far better to power down the transmitter and use a lower value of directional coupling.

Start with the directional coupler in the forward

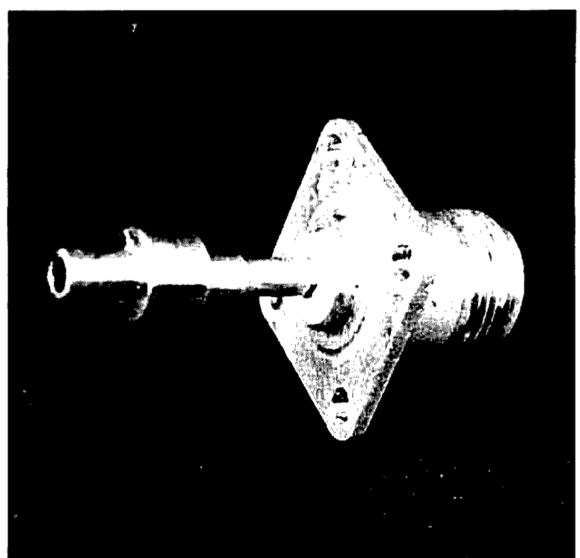


Inside view of unpainted horn showing rf-probe installation.

direction (opposite to that shown in fig. 6), and adjust the voltohmmeter range and the variable resistor so that the meter reading is full scale. Next, turn off the transmitter and reverse the directional coupler; then turn the transmitter on again and observe the vom reading. Adjust the probe length (with transmitter off) for minimum vom reading. Adjust the probe length so that the reflected power is as close to zero as possible. This condition corresponds to minimum vswr.

At this point, switch the vom to a more sensitive scale and adjust the position of the brass washer for minimum vom reading. It should now be possible to reduce the vom reading essentially to zero.

Details of the probe assembly, consisting of a length of brass tubing soldered to a UG-58A/U type N connector. Probe length adjustment and reactance tuning are also provided (see text).



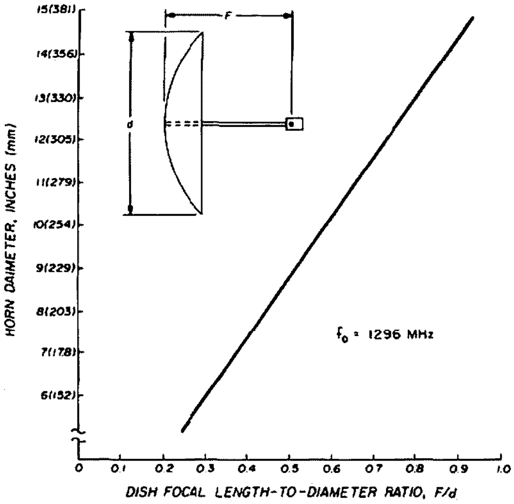


fig. 5. Feed-horn diameter as a function of focal length/diameter ratio, F/d, of the parabolic reflector, for a 1296-MHz system. According to fig. 1, the probe for this horn should be 3.6 inches (91mm) from the shortened end.

An interesting experiment is to hold a metal reflecting plate close to the open end of the horn and note variations in the meter reading. Even at considerable distances from the horn, indications of power being reflected back into the horn can be demonstrated.

The feed horn vswr may change slightly after the feed horn has been mounted on the dish. The reason for this is that a small amount of rf energy reflected from the dish or supporting boom will intercept the horn. Therefore, it's a good idea to use the setup of **fig. 6** to check the reflected power after the horn has been mounted on the dish. This is particularly important for EME installations where even the smallest misadjustments may have serious consequences.

radome

The plastic covers that come with coffee cans make excellent radomes with low loss even at 2304 MHz. These covers fit tight enough to keep the snow and the birds out. As an alternative, a one- or two-inch-thick (25 or 50mm) disc of styrofoam pressed into the open end of the horn also makes a good radome, having a loss about the same as the plastic cover. Unfortunately, birds are attracted to the white styrofoam and chip away at the radome with their bills. A means to solve this problem is to apply coarse fiberglass cloth to the outside of the styrofoam radome with epoxy cement. Don't use polyester resins which react with and dissolve the polyfoam.

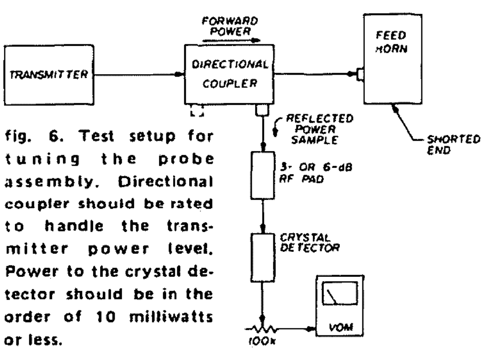


fig. 6. Test setup for tuning the probe assembly. Directional coupler should be rated to handle the transmitter power level. Power to the crystal detector should be in the order of 10 milliwatts or less.

summary

Probably the most important part of the parabolic antenna system design is the feed horn, where small compromises can result in sizeable reductions in overall performance. You can achieve satisfying and rewarding results (and take the guesswork out) by carefully following the design, construction, and tuneup instructions given here. Your parabolic antenna should then operate at relatively high efficiency, an important factor in any antenna system, particularly when conditions are marginal, such as long-haul tropospheric contacts and EME work.

Completed 2304-MHz feed horn ready for installation on dish.

references

1. McKee, Charlton, and Holtum, "Optimizing Gain of Parabolic Antennas," *Microwaves*, March, 1967, page 34.
2. G. Vilardi, WA2VTR, "Simple and Efficient Feed for Parabolic Antennas," *QST*, March, 1973, page 42.
3. G. Southworth, *Principles and Applications of Waveguide Transmission*, D. Van Nostrand Company, Inc., New York, 1950, pages 402-406 and 164-178.

ham radio

six-element collinear antenna for 20 meters

A fixed-azimuth wire array for point-to-point communications

A couple of years ago I had a requirement for a fixed-azimuth array for reception and transmission on 20 meters between Boulder, Colorado, and Europe. A narrow beam was required for receiving European stations at azimuths of 15 to 55 degrees because of strong U.S. East-Coast interference at azimuths near 80 degrees. I was impressed with the narrow-beam characteristics of the collinear array, sometimes called the Franklin antenna.[1-5] The simple collinear, as originally conceived and still described in all the amateur handbooks, is a horizontal wire array. The collinear was later adapted to vhf omnidirectional broadcasting by positioning it vertically and using cylindrical conductors.

antenna description

The array described here consists of two or more horizontal doublets in series. The currents are in phase in each doublet. Ordinarily if half wavelengths of wire are connected directly, the current shifts phase 180 degrees from one wire to the next, so a phase-reversing circuit is used between each doublet. The circuit traditionally used is a quarter-wave open-wire line. Alternatively, a high-Q resonant L-C trap could be used and would have the advantages of small size and light weight. However, it would require high-voltage capacitors and weather protection and might not be as stable in adjustment as an open-wire line.

beamwidth and feed impedance

Reference 6 gives the horizontal beamwidths of collinear arrays. A two-element array has a beamwidth at the half-power points of about 48 degrees; a three-element array has a beamwidth of about 36 degrees. An average three-element Yagi has a horizontal beamwidth on the order of 60 degrees at its optimum vertical angle of radiation. The collinear array is bidirectional. The three-element collinear array has the advantage of symmetrical center feed at a current maximum, so this was my first serious starting point after some poor luck with a two-element array. To feed two elements or any even number of elements symmetrically, one must enter at a high-impedance point. Unbalanced coupling and losses may occur if a metal mast is used to support the feed point.

By Dick Silberstein, WØYBF, 3915 Pleasant Ridge Road, Boulder, Colorado 80301

A three-element 20-meter collinear array was constructed with no. 10 AWG (2.6mm) copperweld wire, and the elements were cut to the usual 95 per cent of a half wavelength in free space. The phasing lines were a full quarter wavelength long but were adjustable from a maximum of about 18 feet (5.5m). The impedance of such antennas, two to six elements in length, is said to be about 100 ohms times the number of elements.[7] I decided to match the antenna to 50 ohms by using a quarter-wave matching section according to the well-known relationship

$$Z_o = \sqrt{Z_a Z_b} \quad (1)$$

where Z_o is the characteristic impedance of the matching section, Z_a is the input impedance, and Z_b is the output impedance.

I found experimentally that with an output impedance of 50 ohms, the characteristic impedance of the matching section had to be 140 ohms to match the input impedance. From **eq. 1**, the antenna resonant impedance was about 392 ohms. By the rule given in reference 7 stated above, I'd expected something nearer 300 ohms.

The matching section was made from two quarter-wave sections of old Belden no. 8275 twinlead connected in parallel, but almost any good 300-ohm twinlead would have been satisfactory. The actual impedance of this particular transmission line was 280 ohms. Two sections in parallel obviously gave 140 ohms. The velocity factor came out by test to be 86 per cent instead of the usual 82 per cent, making the correct length 15 feet (4.6m). The antenna swr was near unity between 13900 to 14350 kHz.

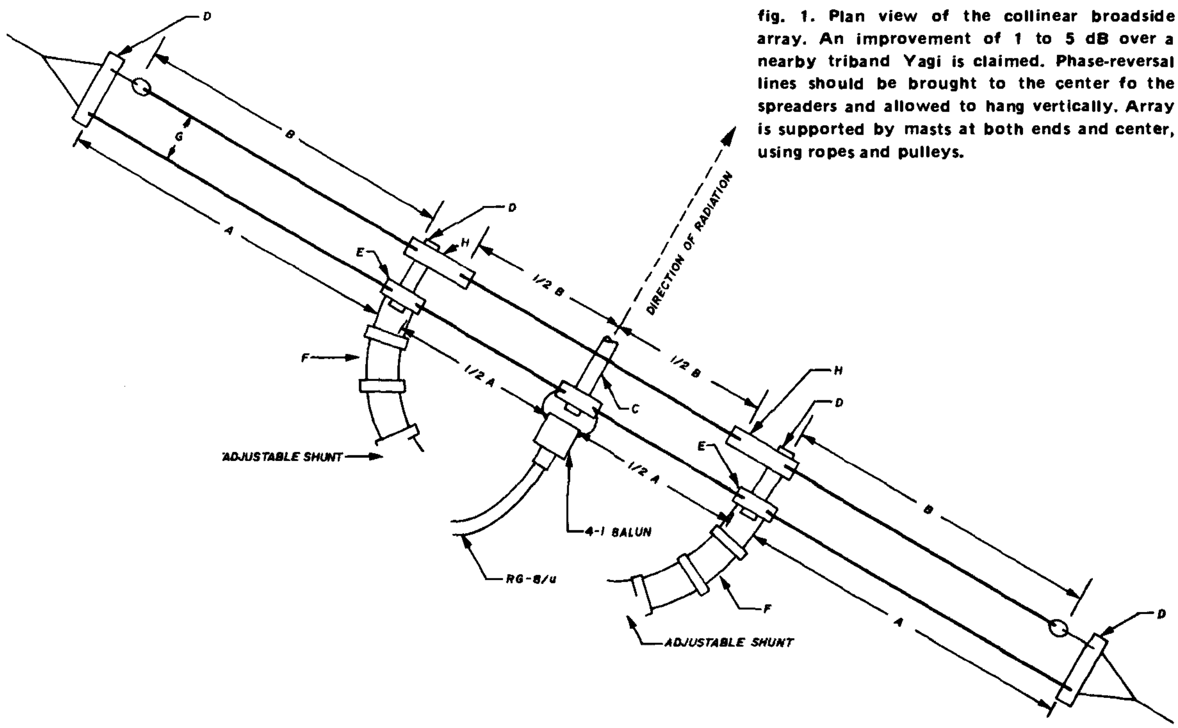


fig. 1. Plan view of the collinear broadside array. An improvement of 1 to 5 dB over a nearby triband Yagi is claimed. Phase-reversal lines should be brought to the center fo the spreaders and allowed to hang vertically. Array is supported by masts at both ends and center, using ropes and pulleys.

A. Driven elements	no. 10 AWG (2.6mm) copperweld wire
B. Directors	no. 10 AWG (2.6mm) copperweld wire
C. Center spreader	1-1/8 inch (29mm) diameter oak pole (8 feet or 2.4m long in 20-meter array)
D. Spreaders (4 required)	1 inch (25mm) PVC tubing (8 feet or 2.4m long in 20-meter array)
E. Insulators	½ inch (12.5mm) nylon or 1 inch (25mm) diameter PVC tubing 6 inch (15.2cm) long
F. Phase-reversal lines	see text
G. Separation distance	0.1 wavelength
H. Insulators	1 inch (25mm) diameter PVC tubing (2 feet or 61 cm long in 20-meter array)

During test, this antenna produced slightly stronger signals from Europe than my three-element tribander. However, to eliminate bidirectionality, increase gain, and narrow the beam a little bit more I decided to proceed further.

The most obvious approach was to make the collinear a driven element of a two-element parasitic array. The only place I've seen anything like this is in reference 6, where mention is made of a reflector with a collinear array. However, from a mechanical standpoint, a row of three parasitic reflectors would be difficult to construct since it would have to be longer than the driven array,

requiring short stubs to make it fit parallel to the driven array. I chose instead a row of three directors, which would be shorter than the driven-array assembly. Each director was separated from its neighbor by a section of plastic tubing.

The theoretical lengths for a two-element parasitic antenna were used for each driven element and its correspond-in director, according to reference 6 for two-element arrays and reference 8 for the driven element and director of a three-element array:

$$Driven\ element\ (ft) \cong \frac{475}{f}$$

$$Driven\ element\ (m) = \frac{144.8}{f} \qquad (2)$$

$$Director\ (ft) \approx \frac{455}{f}$$

$$Driven\ element\ (m) = \frac{138.7}{f} \qquad (3)$$

where f = frequency (MHz)

Construction details are shown in **fig. 1**. Note the generous use of rigid PVC water pipe, which provides good mechanical strength and rf insulation in a lightweight material yet doesn't warp enough to affect spacing appreciably. The antenna was built for 20 meters. Nominal values of length for 14.2 MHz are

driven elements 33 ft, 5 in (10.2m)
directors 32 ft (9.8m)
separation 7 ft (2.1m)
phase-reversal stubs 17 ft 4 in. (5.3m)

The phase-reversal stubs were actually 18 feet (5.5m) long with adjustable shunts. These shunts are needed for correcting minor errors in element lengths, compensating for height differences in different installations, and making final swr adjustments.

The antenna is driven through a 4:1 balun at the end of an RG-8/U feedline for the following reasons: The resonant impedance of a simple three-element collinear measured experimentally in the same location was almost 400 ohms. From experiments with parasitic antennas I reasoned that good gain was possible with element adjustments, which would reduce the normal resonant impedance of an antenna having no parasitic elements by a factor of about one-third. This means that the 400-ohm impedance of three collinear elements alone would be reduced to about 133 ohms with directors under certain adjustment conditions. A 4:1 balun would make the input impedance about 33 ohms, for an swr of approximately 1.5.

swr measurements

Swr was adjusted for a minimum at the desired operating frequency by moving the shunts on the phase-reversal stubs. The swr varied between 1.5 and 2.0 over most of the 20-meter band. Improved results might have been obtained with a gamma match, but I didn't use one because the match might not have held over a wide enough frequency range. Certainly it isn't good to push stub lengths too far beyond their theoretical values in optimizing swr alone. This is especially true in view of information in reference 8, which shows only 14 ohms impedance for a two-element parasitic array at 0.1 wavelength spacing, adjusted for optimum gain. The free-space impedance of the driven element alone would be 72 ohms (at resonance), giving an impedance reduction of about 5:1 due to the presence of the parasitic element, rather than the 3:1 ratio assumed above. So it would be a worthwhile experiment to start with theoretical values of element and stub length, adjust swr on a gamma match, and make corrections of driven-element length by moving the shorting stubs.

The antenna as built had some characteristics of an inverted vee. The antenna center was suspended 25 feet (7.6m) above ground, but the ends were only about 10 feet (3m) above ground on one side and 15 feet (4.6m) on the other. Proximity to ground increases losses but doesn't affect horizontal beamwidth.

results

Performance tests over about 18 months showed that the six-element array in reception gave S-meter readings on most European signals 1 to 5 dB stronger than those received on an adjacent triband Yagi. Interference from East-Coast U.S. stations was about 8 dB below that noted on the tribander, yielding even greater net signal-to-interference ratios; the advantage frequently was two S-units or more.

final remarks

The main disadvantage of this antenna is the same as any other antenna system using resonant phasing stubs or feedlines: the antenna impedance changes greatly under icing conditions. However, here in Colorado, such conditions exist only for a few hours each winter. The six-element broadside wire array should provide excellent narrow-beam, fixed-azimuth reception and transmission even at low heights. The 20-meter version is fairly long, yet it should fit into a half acre (2024m^2) of land. It is an easy-to-erect, high-performance antenna for those wishing point-to-point communication with a distant station.

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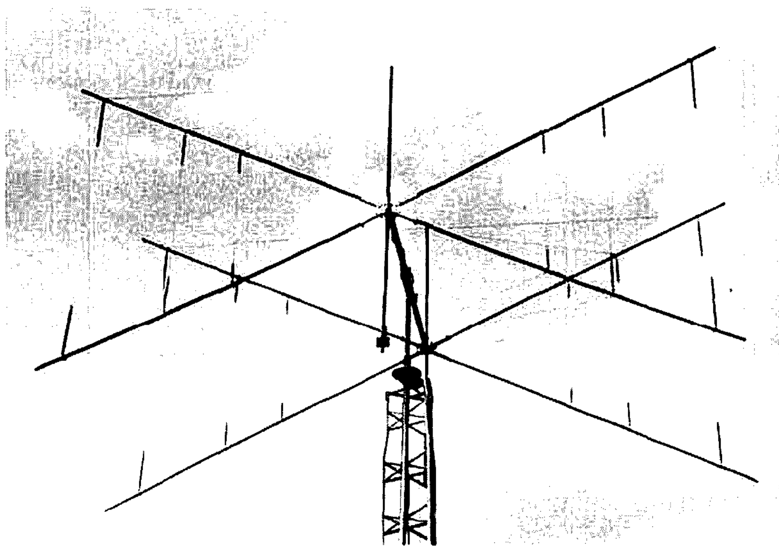
ham radio



improved low-profile three-band quad

An updated version of this compact antenna which features higher structural strength and a different tuning method

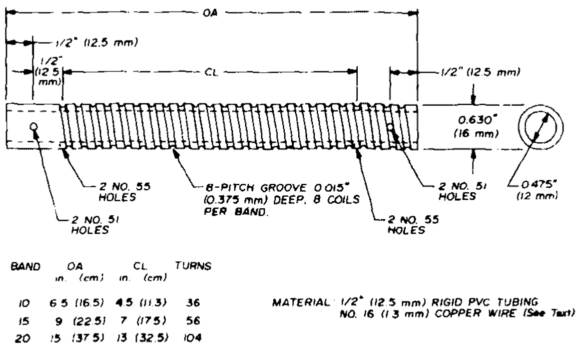
My previous article on the low-profile quad[1] ended with a statement about a new model with better structural rigidity. The version described here has been designed with hexagonal-shaped 10- and 15-meter elements, providing such improvement. Another innovation has been added which consists of loading coils in place of the folded three-wire section and its shorting bars. This doesn't imply that the loading-coil version is better; it's merely another method resulting in easier element assembly. The biggest problem is making 24 loading coils, 8 for each band. I've provided instructions for making these coils using simple shop equipment.

The low-profile quad is unique[2] because the basic quarter-wavelength radiator sections have been retained and compactness achieved by tampering with the "no-good" quarter-wavelength vertical antenna. It's surprising to find that many quad users are unaware that only 50 per cent of a full-wavelength quad is being used effectively. The vertical sections are out of phase, and radiation fields cancel. The primary purpose of the ver-

By John P. Tyskewicz, W1HXU, 77 West Euclid Street, Hartford, Connecticut 06112

tical sections is to complete the full-wave loop. If a small loss occurs because of the closer spacing, such loss is compensated by a smaller and stronger structure.

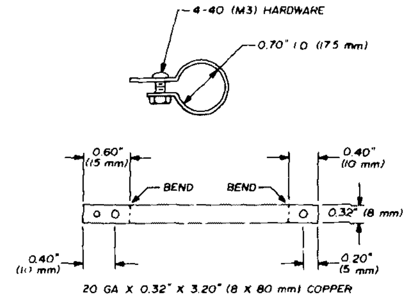
The loading-coil shape factor isn't critical. *I used ½ inch (25mm) (nominal) PVC plastic tubing because of its low wind resistance. Some builders may be tempted to use one large coil at the center of each vertical leg; however, this wasn't tried because the high rf voltage at this*



LOADING COIL FORM

point could be a trouble spot. Wire size isn't critical because the coils are located 1/8 wavelength or more from the high-current points. I used no. 16 AWG (1.3mm) copper wire salvaged from a discarded electric motor. Coils are exposed to the elements, but no serious detuning occured during rain or snow. Fig. 1 *shows the* sets of coils required. To avoid any mixup, driven-element and reflector coils are identical.

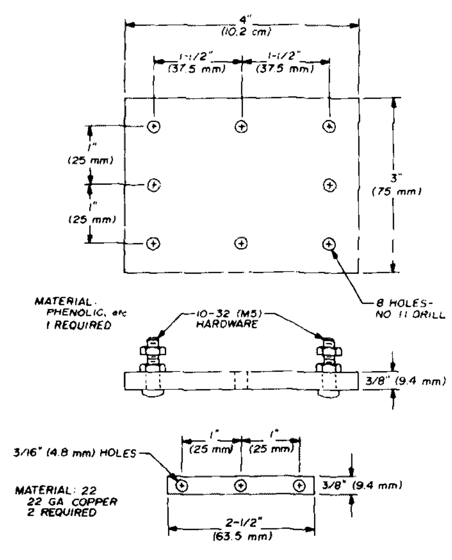
The **coil forms** are made by machining a groove on the PVC tube. This was done on a lathe set up for an 8-pitch thread. A steel mandrel was made to fit the inside diameter of the PVC tube. One-half inch (25mm) of the tube *was held* by the lathe chuck. The cutting tool was ground to a radius slightly larger than the wire diameter. The cut was made in two passes: the first was 0.010 inch (0.25mm) deep, followed by a second cut 0.005 inch (0.13mm) deep.



SHORTING SLIDER CLAMP

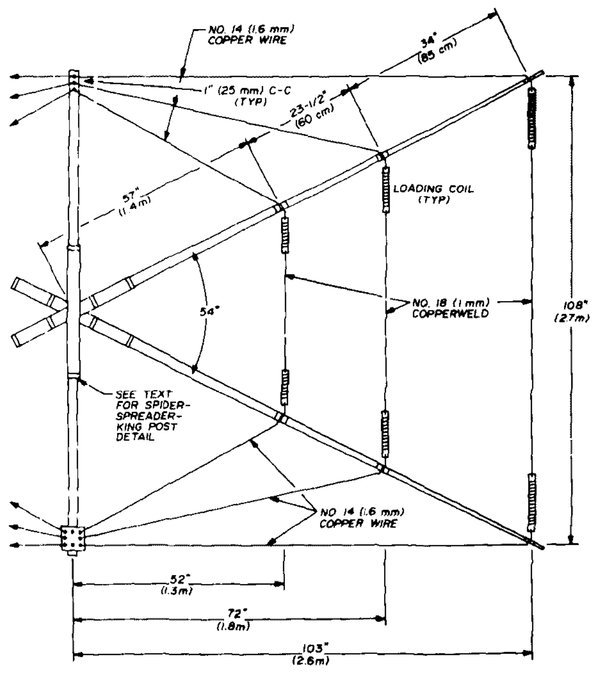
fig. 1. Construction details for the coil forms and shorting clamp.

The coils are wound by stretching a length of wire held in a bench vise and rotating the coil form by hand to accept the wire. Gloves are a must. Wrap two layers of masking tape around the end of the coil, cut the wire to allow a 6-inch (15cm) lead, and pull the wire through *the holes in the form. Insulation should be cleaned from* the bottom end of each coil for several turns (see fig. 3) so that an adjustable clamp can be moved over the coil



LOOP CONNECTORS

fig. 2. Details for constructing driven element and terminal block (1 required).



for tuning. Only one set of upper and lower coils per element need be tuned.

element assembly

Construction of spiders, spreaders, wire anchors and boom was covered in reference 1. **Figs. 2** and **3** show the radial dimensions necessary to give the proper spans. The final dimensions can vary a bit, preferably on the long side, to ensure sufficient tuning range. Fastened to the bottom of the driver element king post is an insulated terminal block, **fig. 2**.

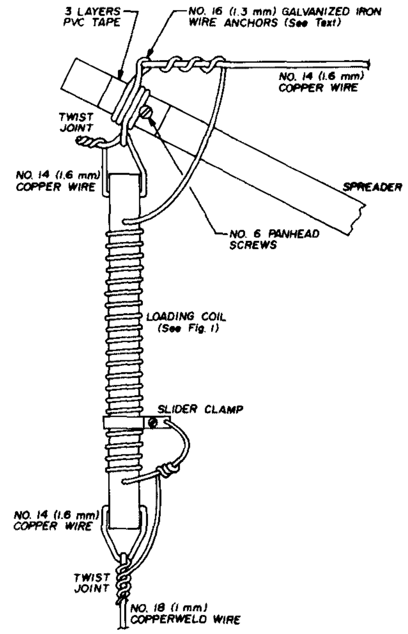
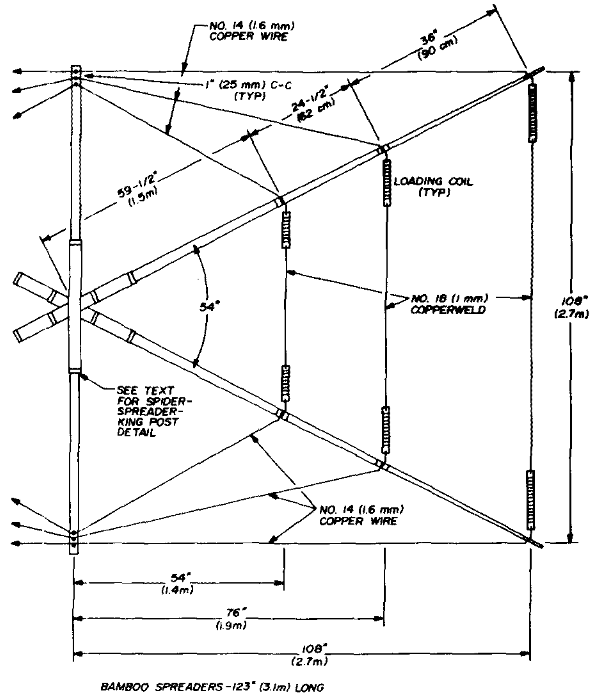


fig. 3. Reflector-element construction and loading-coil installation.



Wire stringing is started with the 20-meter loading coils and wire pieces, followed by the upper and lower horizontal wires. Next, the 15- and 10-meter sloping wires are added, and finally the remaining coils and wire parts. The coil leads are connected to the driver and reflector wires. *Do not* rely on continuity through the wire anchors!

tuning and adjustment

Initial tuneup can be done near ground level with a grid-dip meter monitored by a calibrated receiver. At the driven element terminal block, close the element loops with an insulated wire jumper having a one-turn loop to which the gdo can be coupled. Scanning will reveal a multitude of dips because of mutual coupling between three resonant elements. The trick is to isolate the correct dip by its second harmonic. With full-turn loading coils the resonant frequency should be at the low end or perhaps out of the band. The frequency is now increased by shorting turns from the two lower or upper coils. If more than half of the two coils must be shorted, then the respective opposite coils should be partially shorted. Do not remove any turns at this time.

The reflector coils can now be adjusted to the same settings as those of the driven-element coils, providing the reflector-element span is approximately 5 per cent longer; otherwise make an educated guess. Recheck the driven-element loops with the gdo. Remove the loop jumpers from the terminal block and install the vertical strap jumpers, connecting the three loops in parallel. A homemade balun[3] was used with the RG-8/U, 50-ohm coax.

Direct feed with a single coax cable is not the best method; nevertheless, it works out quite well. Because of some reactance, the swr is sensitive to line length – a case of conjugate tuning. The resonant frequency of the elevated antenna can be quickly determined with an antenna noise bridge. A simple field-strength meter was used for front-to-back ratio and forward-gain adjustments. Current expert opinion[4] is not to lose any sleep if the swr can't be decreased to less than 2:1. The low-profile quad with loading coils and the three-wire version seem to be equally good; each has its merits.

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ham radio

selective antenna system

for minimizing unwanted signals

An approach to the interference problem using matched vertical antennas and a novel phase-control method

Evening operation on 40-meter phone has long been complicated by a wall-to-wall array of foreign broadcasters whose power levels present a formidable barrier. The result, too *frequently*, is amateur inactivity or a mass migration to the already over-crowded 75-meter band.

Isn't it about time to do something more objective? A favorable outcome of the 1979 World Administrative Radio Conference is not assured, so perhaps we should reexamine our techniques and serve notice that we don't intend to be dislodged. Increased power would be barking up the wrong tree, as we are faced with a receiving problem. "You can't work 'em if you can't hear 'em!" A compact, highly directional receiving antenna is required, which is inexpensive and easy to erect.

The best receiving antenna is one giving the best signal-to-noise ratio, not merely the strongest signal. Below 30 MHz, weaker signal pickup can usually be improved by amplification without penalty. The following account is not presented as a solution to the problem, but as a possible starting point for those sufficiently interested in doing something about it.

background

The first effort[1] used two widely spaced verticals in a steerable phased system. Because of the wide spacing between antennas and lack of amplitude-balance control, this system did not produce the deep null required for this application.

In a second effort the phasing system of reference 1 was retained but a pair of very small whip antennas, 2 feet (61cm) high were substituted, each with its own transistor preamplifier. In addition to providing necessary gain, these preamplifiers isolated the antennas from effects resulting from phasing-control adjustments. This allowed closer antenna spacing to reduce diversity reception differences between signals obtained from the two antennas. Initial spacing of a quarter wavelength was nonproductive, so a reduction was made to about 6 feet (2 meters).

To control preamplifier gain for amplitude balance, one unit used an npn transistor, while the other was a pnp type. The method of controlling the applied voltage to each unit is apparent from **fig. 1**, where a potentiometer bridges the dc power supply, with the movable arm grounded to the coaxial cable outer conductors.

The null depth first obtained left much to be desired until I found that the coils in the phasor were picking up the signals directly. When these coils were rewound on toroid forms, the situation was greatly improved. Shielding the original coils should have been equally effective.

By Henry S. Keen, W5TRS, Fox, Arkansas 72051

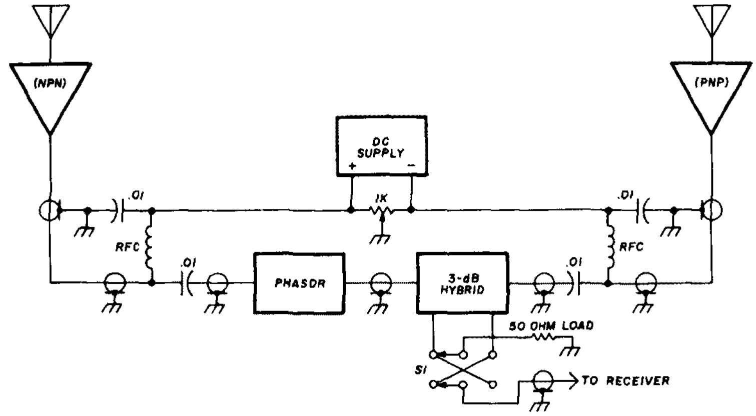

fig. 1. Setup used for receiving tests with foreign broadcast signals on the 40-meter band. Antennas are 2 feet (61cm) long spaced about 6 feet (2m). The hybrid coupling system, described in reference 1, was used to control received signals.

receiving tests

Tests made on the signal of Radio Moscow, in late March, 1975, showed a 50 to 55 dB difference between the in-phase and out-of-phase condition. These numbers, in the absence of calibrated test equipment, were based on an assumed 6-dB per S-meter division reading. This degree of improvement was not all "gravy," however, as the close spacing of the antennas reduced all signals 10 to 15 dB from the in-phase signal level. However there were happy occasions when an otherwise completely smothered signal could be dug out for solid copy.

Later in the year, and through the early summer of 1975, when conditions were particularily poor, the gain achieved over the broadcaster was degraded appreciably, but noticeable and worthwhile improvement was still there. It seemed ironic to find that the very conditions of selective fading and other multipath propagation effects, which degrade normal reception of the broadcast signals, were the same conditions that limited the effectiveness of the phase-out.

The tiny antennas used in this experiment were chosen to approximate the voltage probe antenna,[2] with the hope of making the system less sensitive to the polarization of the incoming wave. The input amplifier mentioned in reference 2, however, used fet elements instead of bipolar transistors, which required a tuned network input circuit. As this investigation was aimed only at the 40-meter band, this point was not considered critical.

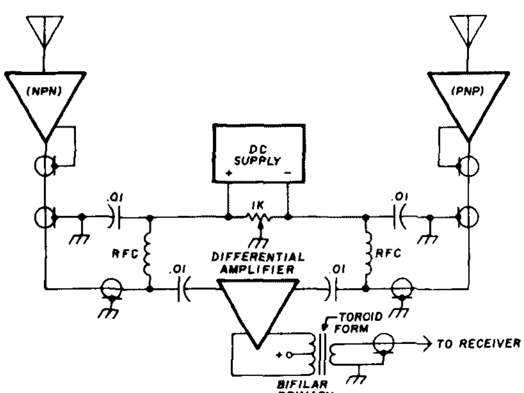

fig. 2. Suggested system using a differential amplifier to replace hybrid couplers of fig. 1. Antennas and preamplifiers would be on a rotatable boom, which could be oriented to minimize interfering signals.

simplified system

A suggested simplified system, which does not require the phasor and hybrid but should be equally effective, is shown in fig. 2. Identical lengths of coaxial cable would bring the signals from the antennas to a differential amplifier, such as the CA3028, with a balanced output. A transformer with a bifilar-wound primary on a toroid core would cancel in-phase signals. The antennas and preamplifiers would be mounted on a rotatable boom oriented to minimize the interfering signal. Amplitude balance of the antenna preamp outputs would be accomplished as before.

With this system multiband operation should be feasible if the preamplifiers as described are replaced with untuned fet input amplifiers as in the voltage probe antenna. This would seem to be an excellent area for further investigation.

Considerable time and experimentation were directed toward different methods of detection that might discriminate against the a-m signal of the broadcaster in favor of the ssb amateur signal. Success in this area was quite limited, however, as multipath propagation problems complicated matters and severely limited benefits that otherwise might have been obtained.

conclusions

The conclusions drawn from the study are that the receiving antenna is by far the most likely candidate for improvement, and that the general concept of using the same antenna for both transmitter and receiver severely limits the ability of the amateur station to compete under conditions of heavy interference.

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ham radio

loop-yagi antennas

Comparative data based on recent literature is presented on loop-Yagis versus conventional designs

The elements of a Yagi antenna can take a variety of shapes. The most common are linear designs, with elements arranged as in fig. 1; the square loop, as in the quad; or triangular loop, as in the delta loop. Less often seen are circular loops. Yagis having an element formed into a loop, usually of one-wavelength circumference, are called loop-Yagis.

The controversy of loop-Yagis versus linear, or conventional Yagis, has raged for many years. Feeding the controversy have been a number of articles making technical comparisons on a practical basis.[1,2] Usually, however, the choice is made on the basis of mechanical convenience, appearance, or plain dollars and cents. I'd like to draw attention to two recent papers that detail design methods for optimum-gain arrays: one for Yagis of conventional design;[3] the other for loop-Yagis.[4]

conventional yagi optimization

"Optimum Element Lengths for Yagi-Uda Arrays," by C. A. Chen and D. K. Cheng,[3] appeared in the January, 1975, issue of the IEEE journal, *Transactions on Antennas and Propagation,* and is the first paper of interest. (See also reference 5). An analytical method is described that begins with a given design; element lengths and spacings are then shuffled several times until the gain is optimized. Chen and Cheng refer to these adjustments as "length-spacing perturbation," The mathematics of the method are somewhat involved but are amenable to computer solution. Perhaps someone with the time and experience could write a program and oblige the amateur fraternity with published data.

Chen and Cheng[3] illustrated their technique by applying it to a six-element Yagi. This antenna consisted of a one-half-wavelength-long driven element, a reflector about 4% longer spaced a quarter-wave behind the driven element, and four directors spaced 0.31 wavelength apart, all 0.43 wavelength long. The gain of this initial array was 8.8 dBd (gain in dB referred to a dipole). The array parameters were then adjusted for maximum gain, using Chen and Cheng's procedures, ending with a gain of 11.25 dBd. In the process, the length increased from 1.49 to 1.69 wavelengths. (But this alone does not fully account for the gain increase of nearly 2.5 dB). That final figure puts the array in the same ballpark as loop Yagis of similar length, according to references 1 and 2. It also exceeds measured gains of published Yagi designs having more elements. The final design and its parameters are illustrated in fig. 1. Element lengths and spacings for various frequencies, which I computed are given in tables 1 and 2. I haven't tried the numbers in practice;

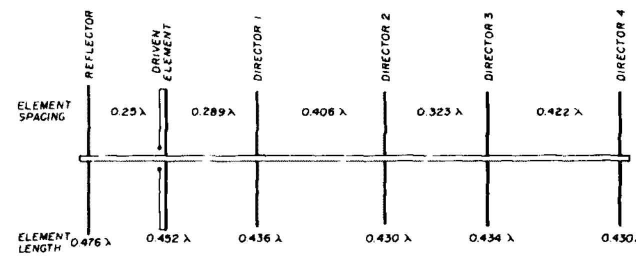

fig. 1. Optimized design of Yagi-Uda array by Chen and Cheng. Drawing is to scale: 1.75 inch (44.5mm) equals 1 wavelength. Element diameter is given as 6.738 x 10^{-3} wavelength.

By Roger Harrison, VK2ZTB, 47 Ballast Point Road, Birchgrove, New South Wales 2041, Australia

they are intended as a starting point for amateur experimentation.

Added benefits gained by the design method of Chen and Cheng are decreased sidelobe amplitude and slightly improved front-to-back ratio (with reference to the initial array). The frontal lobe is narrower as a result of the increased gain. Unfortunately, they make no comment on bandwidth, but bandwidth would be expected to be around 1% or less.

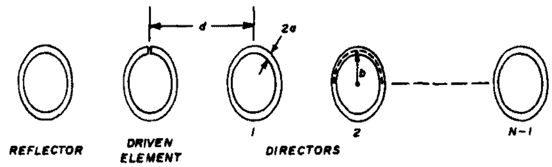

fig. 2. Loop-Yagi antenna (after Shen and Raffoul). Dimensions are loop thickness (*2a*), loop radius (*b*), and loop spacing (*d*).

table 1. Element lengths and diameters for selected frequencies based on the optimized Yagi-Uda antenna of fig. 1.

element	element length (λ)	element spacing (mm except as noted) frequency (MHz) 28.5	50.1	52.1	144.1	146.0	432.1	435.0
reflector	0.476	4.852m	2.760m	2.654m	960	947	320	318
dipole	0.452	4.607m	2.621m	2.520m	911	899	304	302
director 1	0.436	4.444m	2.528m	2.431m	879	868	293	291
director 2	0.430	4.383m	2.493m	2.398m	867	856	289	287
director 3	0.434	4.424m	2.517m	2.420m	875	864	292	290
director 4	0.430	4.383m	2.493m	2.398m	867	856	289	287
element diameter (mm)		69	39	38	14	14	5	5

loop-yagi design

The second paper of interest is "Optimum Design of an Yagi Array of Loops," by L. C. Shen and G. W. Raffoul.[4] They describe a quite simple design procedure. Equal element spacing and element diameter are used throughout. The reflector spacing could be made larger to improve front-to-back ratio (see reference 6).

The general form of the array is shown in **fig. 2**, and one proceeds as follows. The first parameter chosen is usually gain or array size. The curves in **fig. 3** (after Shen and Raffoul[4]) give bandwidth versus array size and gain (in dBd). Select an appropriate d/b (loop spacing/loop radius) ratio or an appropriate bandwidth for the array length chosen. Table **3** (again after Shen and Raffoul) gives the L/λ and b/λ ratios for the d/b ratio just selected. Knowing the wavelength, you can then find b, followed by $2a$ (loop thickness), and thus the distance, d, between the loops. The number of elements (including the reflector) can be found by dividing the approximate boom length by d. The bandwidth decreases with array size (as expected); but even with a large array, the bandwidth is quite substantial.

example

To illustrate the procedure, here's an example. Calculate the wavelength from

$$\lambda = \frac{29050}{f}$$

where λ is wavelength (mm)
f is frequency (MHz)

If f = 433 MHz, wavelength = 671 mm

table 2. Element spacing and array length for selected frequencies based on the optimized Yagi-Uda antenna of fig. 1.

element	element spacing (λ)	element length (mm except as noted) frequency (MHz) 28.5	50.1	42.1	144.1	146.0	432.1	435.0
reflector-dipole	0.250	2.548m	1.450m	1.394m	504	498	168	167
dipole — D1	0.289	2.946m	1.676m	1.611m	583	575	194	193
D1 — D2	0.406	4.138m	2.354m	2.264m	819	808	273	272
D2 — D3	0.323	3.292m	1.873m	1.801m	651	643	217	216
D3 — D4	0.422	4.302m	2.447m	2.353m	851	840	284	283
array length (m)		17.23	9.8	9.43	3.41	3.4	1.14	1.14

From fig. 3, choosing an array length of three wavelengths (3λ), the bandwidth is 13%, or 56 MHz, and the gain is 15 dBd. The a/b ratio is fixed at 0.01. Now, for a d/b ratio of 1.0, proceed as follows.

From **table 3**, b/λ = 0.142, and

loop radius = 0.142 X 671 = 95mm
loop circumference = 2π X radius = 600 mm
loop thickness = $2a$ = 0.02 X 95 = 2mm
loop spacing = 95mm (as d/b = 1.0)
number of elements = $N = \frac{\text{array length}}{d}$ = 21

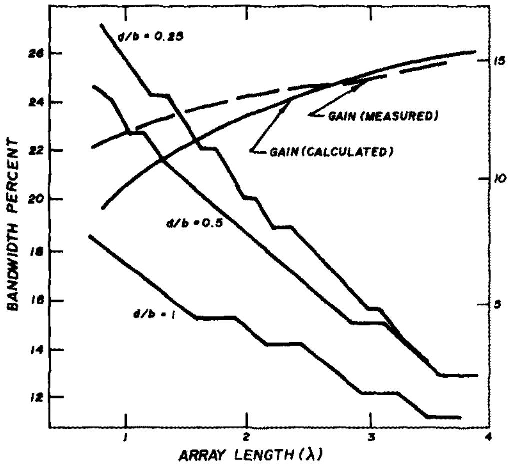

fig. 3. Design curves for loop-Yagi antennas (after Shen and Raffoul). Loop spacing/loop radius values (d/b) are used with table 3 data to obtain array length/wavelength ratio (L/λ) and loop radius/wavelength ratio (b/λ).

Thus the boom length is 2.02 meters. Summarizing,

frequency	=	433 MHz
array length	=	2.02m
gain	=	15 dBd
bandwidth	=	56MHz
loop diameter	=	190 mm
loop circumference	=	600 mm
loop *thickness*	=	*2 mm*
loop spacing	=	95 mm

comparisons

From an examination of fig. 3, a loop-Yagi 1.7 wavelengths long has a calculated gain of 11 to 12 dBd, which compares with the six-element Yagi by Chen and Cheng.[3] But the measured gain of the loop-Yagi is higher than the calculated gain by about 1 dB. Do loop-Yagis still hold the edge in performance? Maybe 1 dB is splitting hairs; a few practical comparison measurements may prove interesting.

table 3. Loop-Yagi antenna design data (after Shen and Raffoul). Ratio a/b is 0.01; L is array length.

d/b = 1.0		d/b = 0.5		d/b = 0.25	
L/λ	b/λ	L/λ	b/λ	L/λ	b/λ
0.73-0.87	0.146	0.78-0.98	0.142	0.81-1.00	0.140
0.88-1.44	0.145	0.99-1.45	0.140	1.01-1.40	0.138
1.45-2.55	0.143	1.41-1.99	0.138	1.41-1.80	0.137
2.56-3.36	0.142	2.00-2.51	0.137	1.81-2.18	0.135
3.37-4.03	0.140	2.52-3.28	0.135	2.19-2.55	0.135
		3.29-3.92	0.134	2.56-3.17	0.132
				3.18-3.65	0.131
				3.66-3.84	0.129

A loop-Yagi 1.7 wavelengths long, designed by Shen and Raffoul's method, has 12 elements. The obvious disadvantage is more hardware than an equivalent-size conventional Yagi; but the wide bandwidth is an advantage, and construction tolerances are relaxed. It would be an interesting exercise to adopt the length-spacing perturbation techniques of Chen and Cheng[3] to the loop Yagi designs of Shen and Raffoul.[4]

table 4. Representative data for an amateur-band loop-Yagi antenna. Dimensions may be used as a starting point for experimentation.

parameter	6 meters		2 meters		70 cm	
gain (dB)	>10	>11	>11	15	>11	14
loop radius (mm)	795	797	288	279	96	95
loop thickness (mm)	16	16	6	6	2	2
element spacing (mm)	795	797	288	279	96	95
number of elements	7	12	12	29	12	21
bandwidth (MHz)	9	7.8	22	16	65	56
loop circumference (mm except as noted)	5m	5.008m	1.81m	1.79m	604	600
physical length (mm except as noted)	5.6m	8.8m	3.2m	7.8m	1056	2013
array length (λ)	1	1.7	1.7	4	1.7	3

construction notes

Dimensions of a representative series of loop Yagi antennas appear in table 4 for various amateur bands. Elements could be made from sheet metal, rod, or tubing providing the loop thickness is maintained; i.e., cross section equal to calculated loop thickness. The elements can be supported by a *metal boom* through the center of the loops, using insulated arms to support the elements. Alternatively, the elements can be supported at voltage nodes (current maxima); i.e. at the feedpoint. Etching the loops on fiberglass PC board would be an ingenious method of construction, although the effect of the fiberglass on the resonant frequency would have to be determined. Insulated boom material, such as PVC conduit, allows elements to be cemented in place using epoxy resin. For further information on loop-Yagis, see references 6, 7, and 8.

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ham radio

towers and rotators

Helpful ideas you can use in planning your rotary beam antenna installation

This article contains some information, not available in current handbooks, that should prove useful in planning your antenna and tower installation, selecting your tower and associated equipment, and raising and lowering the tower. Obviously every situation can't be covered; however, I've tried to include solutions to most problems encountered. Suggestions for preventive maintenance are also provided.

site selection

The average amateur tower is confined to a city lot, so site selection is pretty much determined by nearby structures and available clearance for guys, assuming a guyed tower is contemplated. Much disappointment can be avoided by choosing the safest tower height and antenna load for your location. An excellent treatment on the effect of wind loading on antenna towers in terms of overturning moment appears in reference 1, which is recommended reading for anyone planning a tower installation.

Many cases are on record of antenna towers that had to be removed because of deed restrictions, zoning ordinances and building codes.[2] Check your local ordinances to find out just what restrictions prevail. Assuming there are no problems in this area, the next thing to check is insurance coverage. What can you expect if your tower and antenna end up in the living room of the house next door during a storm?

maintenance considerations

A little planning pays off when you must work on the tower and antenna. It's desirable to perform all maintenance work unassisted, including removal and reinstallation of the rotator. In most installations the antenna is mounted on a mast, which extends several feet above and below the top of the tower and which is mounted in thrust bearings. This mast is driven by a rotator, which is usually mounted inside the tower. A thrust bearing is necessary at the top of the tower to relieve antenna weight and to facilitate certain types of work without having to lower the antenna to the top of the tower. Although a bottom line bearing isn't absolutely necessary, it limits lateral motion at the bottom of the mast, which puts an unequal strain on rotator bearings. When planning the rotator installation, consider the fact that it may be necessary to lift the mast out of the rotator to provide sufficient clearance to tilt the rotator so that it can be passed between the tower braces.

limit switches

After experiencing a tower lift-motor burnout, I installed some limit switches that are activated by the raising and lowering cables. The limits were established by

By E. H. Conklin, K6KA, Box 1, La Canada, California 91011

Crosby clamps, which consist of a U-shaped bolt, a yoke, and two nuts. These clamps were designed to fasten two cables together; installed on one cable, they can operate limit switches.

The switches are standard mercury household types selected with a square-shaped actuating lever to which the clamps and a V-shaped trip wire are mounted (see **fig. 1**). *Leviton* switches have the least taper.

The switches are mounted in standard household metal outlet boxes with cover plates. The boxes are held in place by steel straps. One switch is mounted at the bottom of the downhaul cable (near the tower center), and the other is mounted at the uphaul cable near the cable winch. A bolt at the point of the trip wire secures the cable. A spring between the top of the switch box and the clamp assembly provides tension to keep the switch lever in the *on* position except when it's pushed down by the clamp. Ordinary spring-type clothes pins, clipped on the downhaul cable, can operate the limit switches at intermediate points if desired.

To get the tower out of the limit-switch cutoff position, an interlock override switch is mounted on the motor connection box. This switch shorts the two series limit switches with a momentary contact lever or push-button. Because of heavy motor-starting current, which caused a switch failure after some years of operation, I usually push this button before placing the motor switch in the up or down position and hold it closed until the tower moves out of limit-switch range. I then secure the motor up-down switch to a convenient point to hold it

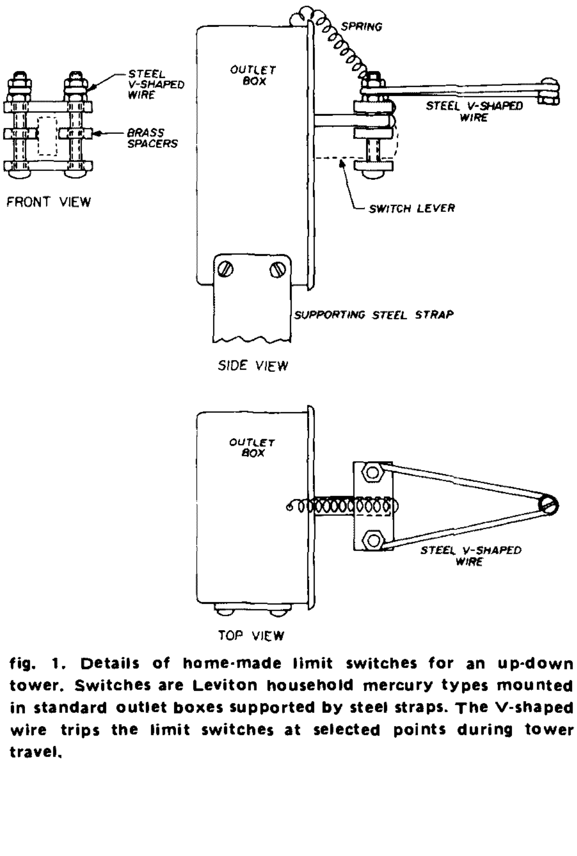


fig. 1. Details of home-made limit switches for an up-down tower. Switches are Leviton household mercury types mounted in standard outlet boxes supported by steel straps. The V-shaped wire trips the limit switches at selected points during tower travel.

on without further attention. Be sure you move the tower in the correct direction to get it out of the limit-switch range to avoid breaking a switch.

guy wires

Guy-wire failure is one of the greatest causes of tower catastrophes. Compromises here will surely result in a mass of twisted metal on yours or your neighbor's

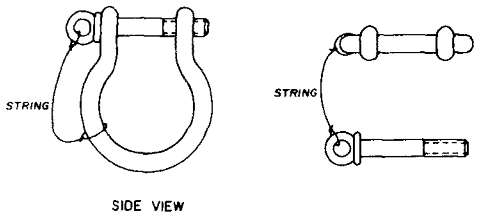


fig. 2. Clevis used to attach snatch blocks to tower braces or heavy eye bolts. These devices are available from marine-equipment supply houses and large hardware stores.

property after a storm. Use the best guy material you can find, with appropriate turnbuckles and locknuts. Check marine hardware suppliers for guywire hardware, such as thimbles and turnbuckles. Never pass an unsupported guy wire through and around a tower brace. Metal thimbles are available for this kind of rigging; they should always be used to relieve friction and eliminate metal fatigue. Crosby clamps should be used in sets of three on each side of strain insulators. Guy-wire anchors should be able to take the tension load of the guys attached to them. Again, reference 1 supplies the geometry and structural information for resolving this problem of antenna-tower installation.

climbing belts

Before you attempt any tower climbing, obtain a new "construction belt." The belt should have a 6-foot (1.8m) tether. Good belts, made of nylon, are available from supply houses serving the construction industry.* Don't buy an old leather belt from second-hand outlets.

The tether might seem a bit long, but it can be snapped around the tower, then tested to determine optimum working length.

antenna assembly and raising

Quads and Yagis seem to be the most popular beam antennas mounted on towers, so we'll discuss these. A quad antenna element can usually be assembled on a driveway or roof. If the quad is completely assembled, it can be placed with its boom next to the tower with the elements straddling the tower. My first quad was mounted on a 20-foot (6m) length of thick-wall aluminum tubing, guyed near the top, and resting on a fitting on top of a chimney. I used a tackle with double-sheave pulleys to lower and raise the mast. I assembled the

*Irving Air Chute Company, Industrial Products Division, Lexington, Kentucky 40500. Ask for model THOR-18.

antenna by standing on top of the chimney, with the quad boom within reach. The boom was slid to one end and an element was attached. I then secured the boom at its center, dressed the cables, and raised the mast without help. A similar approach was used on the tower, assembling one element at a time, with the top mast lowered.

The quad was later replaced by a large Telrex Yagi,

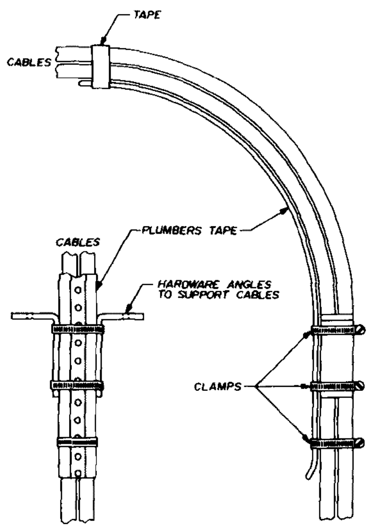


fig. 3. Support assembly for securing coax cable and rotator power lines to the tower. Plumber's tape limits minimum-radius bend for the transmission line, which is important to avoid breakage or short circuits. Clamps are ordinary stainless-steel hose clamps assembled with plastic electrical tape between clamp and cable bundle.

which was assembled on the roof. After assembling the antenna, I raised it to the tower top mast with the antenna elements oriented horizontally. An assembled antenna can be raised using a block and tackle suspended from an S-hook at the top of the top mast. Two lines, each twice the height of the mast, passed over a long boom and back to ground, can be held by assistants to keep the boom level while raising the antenna. Thus the rope can be released at one end and the other end pulled down.

rigging

The block and tackle were moved to the inside of the tower after the antenna was raised. A heavy eye bolt was mounted at the lower part of the top mast so that the mast could be raised to final position with the block and tackle. This assembly is also used to lift the top mast out of the rotator and top thrust bearing when rotator removal is necessary.

Except when used with the S-hook at the top of the top mast, the blocks were secured with a clevis, which is a U-shaped device with a threaded pin closing the open end (fig. 2). The pin has a hole in its head to which a light line may be attached. The other end of the line is attached to the U so that the pin will be there when you need it. Two such clevises are used; one is on a cross brace at the top of the tower, and the other secures the lower block to the heavy eyebolt mounted near the bottom of the top mast. With this arrangement, the entire antenna and top mast can be lowered to the top of the tower when the thrust bearing is released, or mast and antenna can be lifted a few inches out of the thrust bearing.

rotators

Rotator information is summarized for Telrex, WØMLY, and Cornell-Dublier, which produces the TR-44 and HAM-M. Little data is available at this writing on the Hy-Gain design since they no longer market a rotator for amateur use.

Telrex rotators feature worm and chain-link drive, antenna locking, and rotation limit switches. Two minutes are required for full rotation. A mast clamp permits rotator removal, but its size and shape aren't compatible for use with small towers. Rotating torques are available between 6000 and 18000 inch-pounds (69 to 207 kg-meters). Weight ranges between 52 and 145 pounds (23.6 - 65.8kg). Prices are in the $450 to $1100 range. A special 12-conductor cable is required.

The WØMLY rotator requires a 9½-inch (24cm)

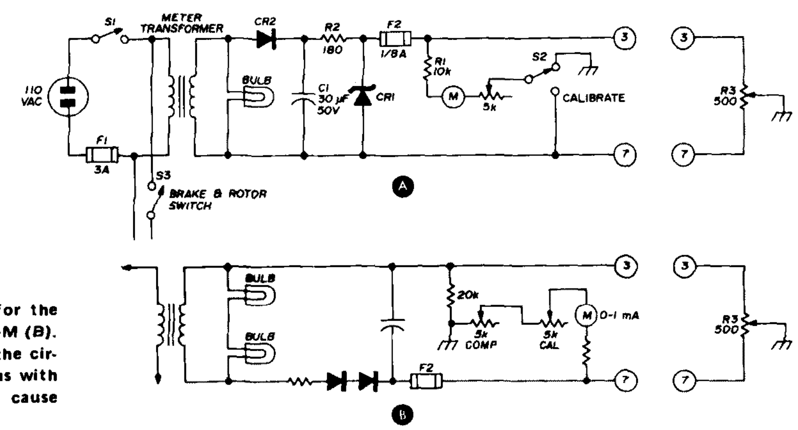
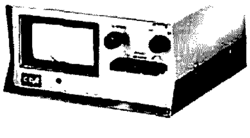
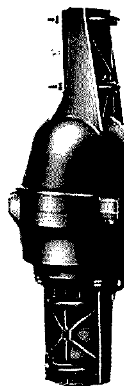


fig. 4. Rotator control circuits for the HAM-2 rotator (A) and the HAM-M (B). The HAM-M can be modified to the circuit of the HAM-2 for installations with varying line voltage, which will cause antenna-bearing errors.

mounting hole. It has a short steel mast to fit inside a 2-inch (5cm) mast diameter. The vertical length of the rotator assembly is about 24 inches (60cm), so it probably won't pass through the side of most amateur towers. It weighs 70 pounds (32kg), rotates in one minute, has high torque, promises 2-degree accuracy from its selsyn system, and its control unit is provided with a





New HAM-2 rotator and control box. Latest model of the popular Cornell-Dubilier HAM series, this antenna rotator system features a new break release control, separate directional control switches, and stainless-steel gears and hardware. The system is designed for up to 7 square feet ($0.65m^2$) of antenna wind load area.

world map (centered on mid USA). Price class is $425.00. A 10-wire cable is required.

Because of extensive experience with the HAM-M at my station, this rotator is covered in more detail below. The TR-44 uses the same control unit as the HAM-M and HAM-2. However, the TR-44 doesn't have a brake solenoid and has a lower rating in terms of antenna wind-area loading, turning torque, and brake torque. The TR-44 has a disc brake, whereas the HAM-type rotators use a steel wedge. The HAM types require not more than one ohm in two of the cables. Rotation time is 48 seconds. These rotators take mast sizes up to 2 inches (5cm) and are 8 inches (20cm) in diameter.

cable support

Cable installation will be determined by your particular situation. However, I'd like to offer some suggestions based on my experience. Coax cable should be dressed so that the radius of curvature is as large as possible. The rotator power cable and coax can be in *one* bundle. Make sure that adequate strain relief and slack are used to avoid problems when rotating the antenna or when raising or lowering an up-down tower. I use the method shown in fig. 3 to support coax and rotator power cables on my tower. A piece of plumber's tape bent into a large radius supports the cable bundle. Two galvanized steel angles keep the assembly from sliding through the hanger on the cable outrigger. Stainless-steel hose clamps tie the assembly together. Small wooden dowels inserted into the cable bundle fill space between cables.

On the lower outriggers, the cable bundle should be tied to one side, which forces the loops to fall beyond the ends of the outriggers when the tower is fully extended. This precaution will prevent a loop of cable from catching around the end of a lower outrigger, which will break either the outrigger or a coax cable. Nylon rope is recommended for cable ties.

control unit mod

A slight modification can be made to the HAM-M rotator control unit that will benefit those with variable line voltage, which can cause considerable error in bearing indication. The HAM-M rotator may be modified by a slight rearrangement of the control-circuit resistors and by installing a 13-volt zener (part no. 50153-00 in the HAM-2 owner's manual). These devices can be obtained from other sources. The zener furnishes 13 volts in the HAM-2 compared with 21 volts in the HAM-M, so the fixed resistors in the HAM-M must be reduced in value and the ground connections changed. Both circuits are shown in fig. 4. The HAM-M circuit is a bridge, whereas that for the HAM-2 uses a simple variable shunt resistance across the meter and multiplier resistors. Either circuit will work with protective 100-ohm resistors installed at the ends of the rotator potentiometer, if the meter zero is reset below zero with current off and the calibration pot is then set for maximum meter reading.

It's well to enter information in your maintenance log such as all resistances between *control unit* and *rotator*, wire color code and connections, and circuit mods as described above. This might help to isolate problems in case of trouble without having to remove the rotator.

If the antenna doesn't rotate when you activate the control unit, check the ac electrolytic capacitor. In addition to the test suggested in the manual, try using two ordinary electrolytics about twice the value of the ac capacitor, connected back-to-back.

factory repair

The HAM-2 owner's manual mentions factory overhaul of a rotator for $15, a control unit for $12, or both for $25. If you need pinion gears or other expensive parts, this is quite a bargain. At this writing, a replacement rotator can be purchased for $82.95 directly from the manufacturer.*

As a final note, I suggest you list *in* your antenna maintenance log every bolt, nut, and other part used in the installation. Another item to include is the wrench sizes for every piece of hardware and where used so you won't be missing some tool when you climb the tower for maintenance.

*Cornell-Dubilier, Rotator Service Department, Fuquay-Varina, North Carolina 27526.

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ham radio

understanding the ZL Special antenna

One answer to the problem of building small, lightweight directional antennas on small parcels of real estate is the ZL Special, a close-spaced version of the two-element driven array. The ZL Special has been around for a long time, but not much has been published about it except for empirically derived data. The ZL Special offers light weight and compact physical size with little compromise in forward gain, front-to-back ratio, or sidelobe levels.

description

The ZL Special basic configuration is shown in **fig. 1**. Two folded dipoles spaced one-quarter wavelength apart are driven 90 degrees out of phase. Typical characteristics are: forward gain, about 3 dB and front-to-back ratio, about 20 dB. Several sidelobes appear when the antenna is placed at heights greater than one-half wavelength above ground. Approximate dimensions are given below, in which F is frequency in MHz, L is element in length, S is element spacing, and P is the phasing-line length for 90 electrical degrees of phase difference between elements:

$$L, \text{ element length (feet)} = \frac{468}{F}$$

$$\text{(meters)} = \frac{143}{F} \qquad (1)$$

$$S, \text{ element spacing (feet)} = \frac{245}{F}$$

$$\text{(meters)} = \frac{74}{F} \qquad (2)$$

$$P, \text{ phasing line length (feet)} = \frac{196}{F}$$

$$\text{(meters)} = \frac{60}{F} \qquad (3)$$

In previous descriptions[1] the ZL Special is shown as six tubular pieces comprising two radiating elements driven 135 degrees out of phase. Spacing between elements is on the order of 1/8 wavelength, and a transposed 300-ohm line is used as a phasing section, **fig. 2**. Claims have been made that the feedpoint impedance is about 70 ohms with this arrangement and that the antenna can be fed with 72-ohm line, although this is probably true only in special cases. The design will work, however, and the dimensions usually given are:

fig. 2 dimension, feet	(meters)
A = $438/F$	$134/F$
B = $447/F$	$136/F$
C = $101/F$	$31/F$
D = $122/F$	$37/F$
E = $110/F$	$34/F$

design for optimum performance

A more modern design would use 300-ohm line throughout, with bamboo or fiberglas supports and a simple *aluminum boom*. However, in this case the phasing line will be physically a bit shorter than the desired element spacing. As shown in **fig. 3** maximum gain for a parasitic element will occur at about 0.11 wavelength for a director and 0.15 wavelength for a reflector. Since the ZL Special has a "driven director-reflector," you might expect that optimum forward gain would occur between 0.11 and 0.15-wavelength spacing. This is indeed the case, and maximum gain occurs at about 0.123 wavelength spacing. In no event should less than 0.1-wavelength spacing be used, because not only does gain drop rapidly but the characteristic (feed) impedance changes drastically.

Empirical designs using 300-ohm line have shown that director lengths of $447.3/F$ in feet ($136.3/F$ in meters) and reflector lengths of $475.7/F$ in feet ($145.0/F$ in meters) are nearly optimum. These dimensions are somewhat longer than those given for the tubing version, primarily due to the much narrower width dimension of

By Gary Blake Jordan, WA6TKT, 1012 Olmo, San Jose, California 95129

the 300-ohm elements. For example, in free space one-half wavelength in feet is *492/F* (*150/F* in meters), whereas in practice a folded dipole at ordinary heights will resonate at *468/F* in feet (*143/F* in meters).

Using 14.2 MHz as a design example, a free-space half wavelength is *492/F* = 34.65 feet, or *150/F* = 10.6 meters, which is 5.19 electrical degrees/foot (17 degrees/meter). The ZL Special dimensions are then:

director *447.3/F* = 31.5 feet
(*136.3/F* = 9.6 meters)
reflector *475.7/F* = 33.5 feet
(*145.0/F* = 10.2 meters)

element spacing 0.12 wavelength
= 8.5 feet (2.6 meters)

Compared to a resonant dipole, the director is shortened by (468 - 447.3)/468 = 0.044 or about 4.4%. Similarily, the reflector is lengthened over the dipole by (475.7 -

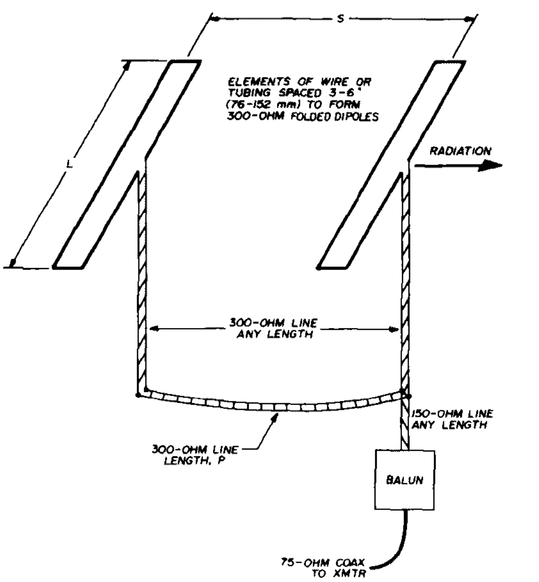

fig. 1. Basic arrangement of the ZL Special, a unidirectional quadraphased two-element array. (Essentially two half-wave antennas phased at 180 degrees.)

468)/468 = 0.0165 or about 1.7%. While these numbers aren't sacred, the difference between them is very close to optimum at (475.7 - 447.3)/447.3, or about 6.3%.

Similarly, for 20 meters a phasing-line length of 7.75 feet (2.4 meters) nearly always proves to be optimum. Making the assumption that the velocity factor of typical 300-ohm line will approximate 0.7, a half wavelength of 300-ohm line is (492/*F*)(0.7) = 24.3 feet, or (150/*F*)(0.7) = 7.4 meters. Then 180 degrees divided by 24.3 yields 7.4 electrical degrees per foot (24.3 degrees per meter) in 300-ohm line. Also, 7.75 feet (2.4 meters) of phasing line yields 57.5 degrees of phase shift.

Since the phasing line transposition adds 180 degrees in phase, the difference in phase between director and reflector is 360 - (180 + 57.5) = 122.5 degrees. Thus in

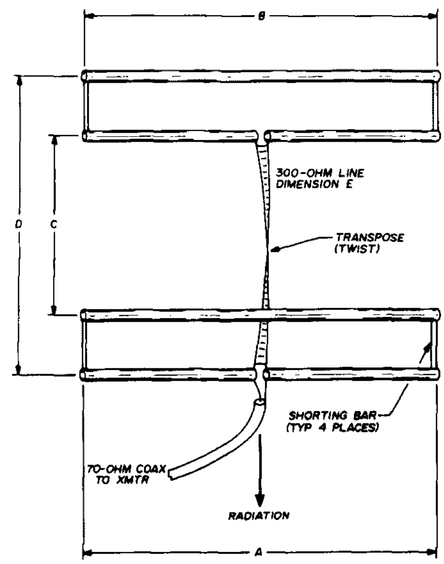

fig. 2. ZL Special using tubular elements. The two radiating elements are said to be driven 135 degrees out of phase; however this value is more like 115-125 degrees (see text).

truth, most ZL Special antennas don't employ 135-degree phasing but rather something between 115 and 125 degrees, depending on phasing-line velocity factor and empirical pruning.

construction

Construction may be as previously described, or as I prefer, using ordinary plastic plumbing pipe (known as PVC tubing). Placing the 300-ohm line into the pipe (no twists allowed) is easy, and T connectors provide additional rigidity for guying (fig. 4). Fig. 5 shows a successful design at 14.2 MHz using the desired phasing line, but with the rear element bowed somewhat to allow for correct element spacing. You may think that the "delta" of the rear element aids in a smooth phase transition (as

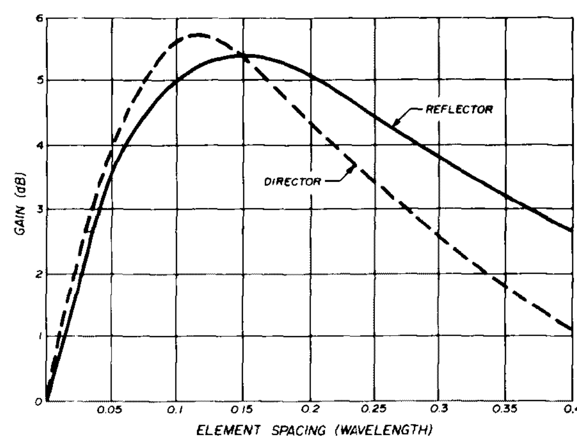

fig. 3. Maximum gain obtainable with a parasitic element over a ½-wavelength antenna alone, assuming parasitic element tuned for maximum gain at each spacing.

with a delta match), but this is a pure speculation. Construction is easy enough with bamboo or fiberglas supports, and still not too difficult with plastic plumbing pipe into which slots are cut to allow the rear element to pass forward to the phasing line.

Rather than simply feeding this balanced antenna with unbalanced coaxial cable, a balanced feed should be used. One method (other than using a balun transformer) is to make a 1/4-wave bazooka line as shown in **fig. 6**. Simply wrap aluminum foil around the last 1/4 wavelength of feed line, using plenty of overlap, then use masking tape to cover the foil. Apply several coats of

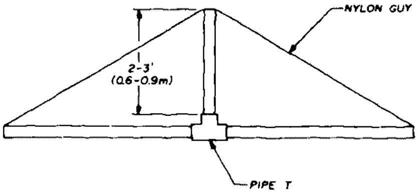

fig. 4. Suggested construction for a single element using PVC tubing and T connector (end view). The 300-ohm line feeds easily into the tubing (no twists permitted).

weatherproof compound to the tape. The foil may be secured to the coax shield by wrapping it tightly with a number of turns of wire.

performance

Performance should be quite broadband compared with a true parasitic beam, and the turning radius for the 20-meter example here will be only 17.3 feet (5.3m). Weight may be less than 10 pounds (4.5kg). Gain over a reference dipole should be 6 to 7 dB, with a front-to-back ratio of at least 15 to 18 dB. Don't forget to take into account the velocity factor of the coax when constructing the bazooka. The 1/4-wave bazooka length is about 11 feet 5 inches (3.5m) at 14.2 MHz. For those amateurs with more space, additional true parasitic elements may be added as in **fig. 7**, although the feedpoint

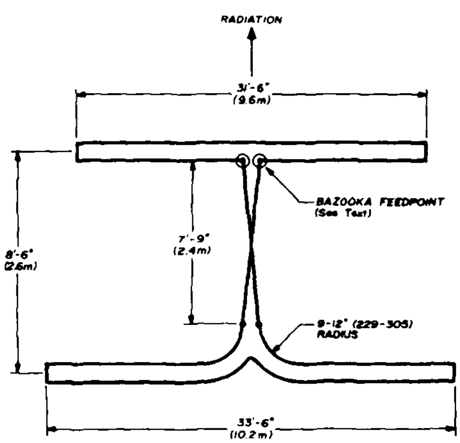

fig. 5. Top view of the ZL Special using 300-ohm twin lead. Design is for 14.2 MHz. Rear element is bowed slightly to allow for the desired element spacing.

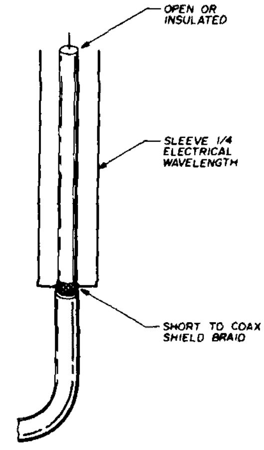

fig. 6. One-quarter wavelength balanced-to-unbalanced transformer for feeding the ZL Special with coax transmission line. Transformer is recommended for keeping antenna currents off coax, which degrade antenna pattern and may cause difficulty in transmitter tuning.

impedance will be lowered. For the basic ZL Special, feeding with 52-ohm line may require that the bazooka be made of 72-ohm line, which will yield a transformation of $(75)^2/52$ = 108 ohms to the antenna. This may be very useful, as the nominal 60 to 80-ohm feed-point

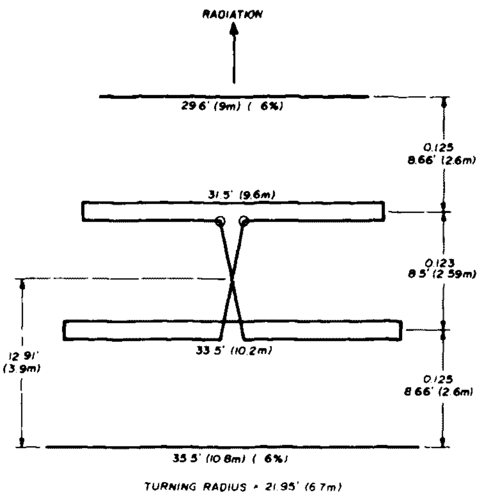

fig. 7. ZL Special antenna with parasitic elements. Typical parameters: input impedance approximately 40 ohms; gain referenced to a dipole at the same height about 13.6 dB; front-to-back ratio 28 to 35 dB. Space for slightly more turning radius is required for this version.

impedance might increase* for small heights (less than one wavelength) above ground. A 52-ohm line plus bazooka will match the ZL Special with parasitic elements reasonably well without further transformation.

The addition of true parasitic elements, when carefully tuned (not an easy chore), can yield gain and front-to-back ratios comparable with parasitic beams which have a greater number of elements.

*On the other hand, the reverse may be true.

reference

1. *The ARRL Antenna Book*, ARRL, Newington, Connecticut, ninth edition, 1960, page 214.

ham radio

5/8-wavelength vertical antenna for mobile work

Problems with loading coils are eliminated with this design — the coax feedline also acts as a matching stub

Most published 5/8-wavelength vertical antennas have used a base loading coil.[1,2,3] I built several of these but difficulty in obtaining components, weatherproofing, and adjusting the antenna for low vswr led me to seek a better design. This design[4] is mechanically simple, uses readily available components, and best of all is easy to adjust for a low vswr over the entire 2-meter band.

The antenna consists of a 5/8-wavelength radiator fed with a length of coax that also is *the matching stub.* A diagram appears in **fig. 1**. The mechanical components are simple. A short length of RG-58/U coax cable with the outer insulation removed and one end shorted, is slipped inside a piece of 1/4-inch (6mm) diameter tubing. The stub is connected electrically in series between the radiator and coax center conductor. The tubing is mounted in an insulator that attaches to a PL-259 coax plug. The feasibility of this design can be demonstrated by making an "emergency" antenna from a 48-inch (122cm) length of RG-58/U or RG-8/U cable, as shown in fig. 2.

electrical performance

A 5/8-wavelength radiator above a ground plane exhibits an impedance of approximately 50-j185 ohms[5] (see **fig. 3** or **table 1**). Thus its resistive component closely matches 50-ohm coax, but it's highly capacitive. To resonate this 5/8-wavelength radiator and provide a purely resistive load, an inductive reactance of approximately 185 ohms is needed, and a loading coil is usually used. A length of coax cable shorted at one end and less than 1/4-wavelength long also appears as an inductive reactance. If a 0.21-wavelength shorted coaxial stub is connected in series with the 5/8-wavelength radiator, capacitive reactance will be cancelled and a 50-ohm resistive load will be presented to the transmission line.

This coaxial matching scheme can be used with many vertical antennas. In the form presented, it can only compensate for an inductive or capacitive reactance.

By Joe Pentecost, K4LPQ, Georgia Institute of Technology, Atlanta, Georgia 30332

table 1. Impedance of radiators mounted above a ground plane with 50-ohm coax feed (calculated from reference 5).

radiator length (λ)	impedance (ohms) 1/4 in. (6mm) diameter	1/4 in. (13mm) diameter
9/16	111-j310	86-j240
19/32	71-j244	58-j195
5/8	50-j185	44-j147
21/32	39-j133	37-j105

Fortunately, radiator lengths between 9/16 to 5/8 wavelength have 40 to 65-ohm resistive components, depending on diameter, and can be easily matched by this technique. I've used this method of "hiding" the matching stub on collinear arrays using four 5/8-wavelength radiators[6] and also with collinear 1/2-wavelength radiators, rather than using conventional 1/4-wavelength open-wire stubs. The advantages of the coaxial design include no radiation from the phasing stubs and ease of constructing weatherproof arrays from available materials.

construction

Detailed dimensions of the components for a 5/8-wavelength whip are shown in fig. 4. The components are assembled as follows. Slip the 36-inch (91cm) long, 1/8-inch (3mm) diameter rod 1-1/2 inches (38mm) into the tubing and solder. You'll probably have to insert a soft copper or brass shim or crimp the tubing to make a tight fit. After joining, the radiator should be 47-1/2 inches (121cm) long overall, and the tubing should be unobstructed for at least 11-1/2 inches (29cm). Next, slip the modified PL-259 connector into the insulator. Epoxy-bond the sleeve, center portion and insulator into a single unit. Be sure to seal between the sleeve and insulator so water can't enter that joint.

The antenna can be made of stainless steel. Stainless-steel welding rods as well as stainless-steel tubing are easily obtainable at low cost. A special soldering flux* is necessary for soldering stainless steel. Use care to clean joints and the inside of tubing to prevent corrosion and to ensure a good solder job. An advantage of stainless steel is that its ductility is good. After several mishaps (garage door, bridges, vandals, trees) it was easy to straighten out an S-shaped whip with no degradation in performance.

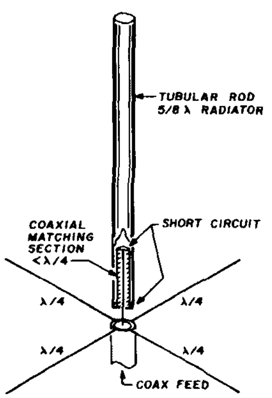

fig. 1. The 5/8-wavelength 2-meter antenna showing series-connected coax matching system. Increasing number of radials will decrease system ohmic resistance and increase radiation resistance.

At this point, you'll have three components: plug and insulator assembly, radiator, and the coaxial matching section. The coax should be carefully soldered at the short circuit so the coax will slide easily into the tubing. Tin the center conductor, cut the end of the braid, and slide the coax into the tubing until the end of the braid is flush with the tubing end. If you wish, the edge of the braid may be carefully soldered to the end of the tubing to ensure better mechanical and electrical stability, although it may also be simply tinned and wedged for a snug mechanical fit. Solder must be kept off the outside of the tubing so that the tubing will slide into the insulator. The coax may be loose inside the tubing with no adverse effects so long as it makes electrical contact near the unshorted end of the braid and can't slip in or out to change its effective length.

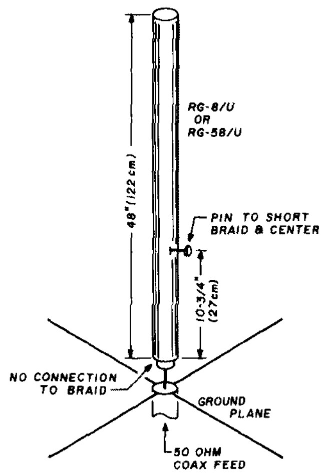

fig. 2. Example of an "emergency" 5/8-wavelength radiator made from coax cable. Outer braid is the radiating element.

When the coax has been inserted into the tubing, measure the distance from the tip of the PL-259 to the top of the insulator. Measure this same distance from the tip of the coax center conductor along the tubing, and scribe the tubing. The radiator tubing should now be inserted into the insulator to the scribe mark and the coax center connector soldered temporarily to check the vswr before applying the epoxy for the final assembly. The assembly shows less than 1.1:1 vswr over the entire *2-meter band. If not, check the dimensions of the* coax and radiator carefully, and be sure braid and tubing are flush in the insulator. A 1/4-inch (6mm) error in the coax length will make a difference in vswr. If you wish to make the overall whip length somewhat shorter, say 42 or 43 inches (107 or 109cm), it will be necessary to

*"Stay Clean" brand flux and "Stay Brite" solder are good for this purpose.

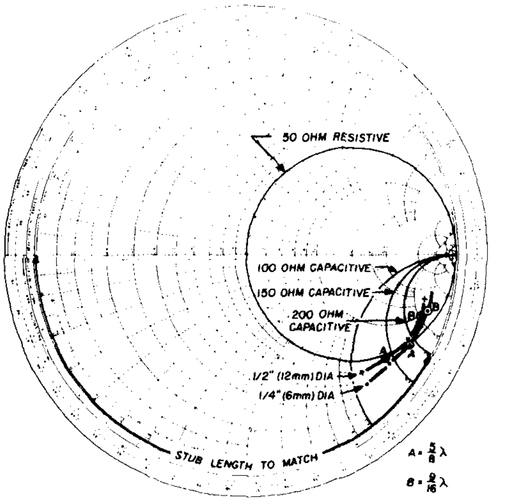

fig. 3. Smith chart showing impedance of 5/8-wavelength radiator mounted above a ground plane and fed with 50-ohm coax cable.

lengthen the coax matching section about 1 inch (25mm).

If the vswr is not very low, check at two frequencies, about 2 MHz apart if possible, and determine which vswr is lower. If the lower frequency shows a lower vswr, shorten the coax or shorten the radiator. If the higher frequency shows the lower vswr, the reverse applies.

vswr measurement notes

Most reflectometers and swr bridges don't appear as a purely resistive 50-ohm length of coax. When inserted into a flat (matched) line they may show an swr *not* representative of the *true line swr*, depending on the line length between bridge and load. When the "impedance" of the vswr meter is placed a multiple of one-half wavelength from the load to be measured, both appear effectively in parallel, sometimes causing questionable results. This is particularly true when very low (less than 2:1) vswr is being measured.

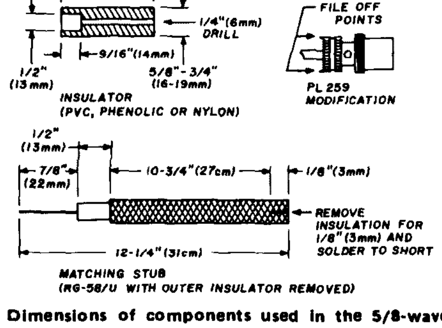

fig. 4. Dimensions of components used in the 5/8-wavelength vertical antenna. Brass or stainless steel may be used for the radiator; the latter is recommended (see text).

After much frustrating *experimentation, I found that* the *best* distance to place a vswr meter from the measured load is an odd multiple of one-quarter wavelength at the *measuring* frequency. Vswr measurements may be checked by adding short 1/8 to 1/4 wavelength lengths of coax to the line between reflectometer and load. For impedance measurements as well as vswr, I use a carefully constructed slotted line.[8] However, such a device is rather impractical to use on a roof or tower.

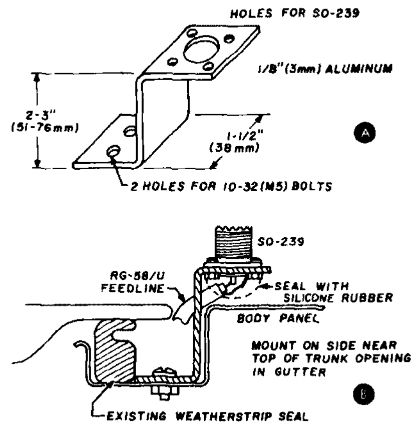

fig. 5. Mounting bracket (A) and suggested mounting details for an automobile trunk lid (B).

For best performance the whip should be mounted on a good ground plane. A mounting for a trunk-lid lip, which requires only two holes (invisible and easily patched), is shown in **fig. 5**. This antenna design can also be used on mounts that use the equivalent of an SO-239 fitting.

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3. Dave Sargent, K6KLO, "5/8-Wavelength Two-Meter Antenna," *ham radio*, July, 1974, page 40.
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6. Bob Dahlquist, WB6KGF, "Four-Element Collinear Array for Two Meters," *ham radio*, May, 1971, page 6.
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ham radio

test data on 1/4- and 5/8-wavelength vertical antennas for two-meter mobile

Measurement results made under practical conditions to help you decide which antenna is best

Controversy still persists on the merits of 1/4- and 5/8-wavelength vertical antennas for two-meter mobile service. Measurements are difficult to make under conditions of multipath transmission between automobile and receiver. The elegant (but impractical) way to make a comparison is to drive over a prescribed route, with each antenna *in turn continuously emitting* constant power, while recording the instantaneous intensity of the signal.

Lacking such means, we decided to make a number of measurements at many fixed points on the road. At each point the 1/4-wavelength and 5/8-wavelength antenna was plugged into the bulkhead fitting on the top of my car. Intensity measurements were made by W2CQH using a precision attenuator on the i-f of his receiver to obtain constant S-meter values.

test conditions

The vehicle was a 1965 Plymouth Fury station wagon. Vehicle overall dimensions are 17.5 feet (5.3m) by 6 feet (1.8m). The roof is 9.5 feet (3m) by 4.6 feet (1.4m). The antenna bulkhead fitting was mounted in a hole drilled through the roof centerline 2.75 feet (0.8m) behind the windshield.

The 5/8-wavelength antenna was 3.96 feet (1.2m) long, fed through a 5½-turn coil, 1 inch (25mm) long and 1/2 inch (12.5mm) inside diameter, made with no. 14 AWG (1.6mm) copper wire. The top part of the antenna was a chrome-plated segmented whip 1/4 inch (7.5mm) in diameter, tapering to 1/16 inch (1.5mm). The reflected power was −19.5 dB.

The 1/4-wavelength antenna was 1.4 feet (0.4m) from the top of the PL-259 connector, which was 1.5 inches (38mm) high. Material was 1/16-inch (1.5mm) diameter stainless steel. Reflected power was −20.5 dB.

Equipment used for the mobile tests was a Midland 13-500 operating at 147.63 MHz. The transmitter delivered 13.4 watts to the bulkhead fitting through a 4-foot (1.2m) length of coax cable.

Transmitting points were 1.2 and 7.7 miles (2 and 12km) from the receiver. At each location the vehicle was moved 3 to 20 feet (0.9 to 6m) in a random fashion.

By Bill King, W2LTJ, and Reed Fisher, W2CQH, 5 Midwood Drive, Florham Park, New Jersey 07932

table 1. Attenuator settings for a constant S-meter reading, test point 1 (1.2 miles or 2 km from receiver).

attenuator setting (dB)		
1/4 λ antenna	5/8 λ antenna	Δ dB
22	27	5
24	26	2
17	20	3
17	23	6
11	16	5
20	22	2
14	20	6
24	26	2
23	26	3
22	24	2

average gain, 5/8 λ over 1/4 λ antenna = 3.6 dB. one standard deviation = 1.7 dB.

Tables 1 and 2 show the attenuator settings at the receiver and the differences between each antenna at each transmitting point. Note that the statistical term "one standard deviation" is used in these tables. In table 1, for example, one standard deviation is shown as 1.7 dB. A standard deviation of 1.7 means that the average is estimated to lie between 3.6 + 1.7 and 3.6 - 1.7 with 67% confidence.*

Remarks from contacts with amateurs who heard about these experiments led to further experiments to compare a stainless-steel and a copper 1/4-wavelength whip under the same conditions. These tests were conducted with no difference in gain noted between the two antennas.

table 2. Attenuator settings for a constant S-meter reading, test point 2 (7.7 miles or 12 km from receiver).

attenuator setting (dB)		
1/4 λ antenna	5/8 λ antenna	Δ dB
12	15	3
11	11	0
10	10	0
-3	4	7
9	15	6
16	18	2
18	18	0
19	20	1
17	17	0
12	13	1

average gain, 5/8 λ over 1/4 λ antenna = 2.0 dB. one standard deviation = 2.6 dB.

concluding remarks

The purpose of this experiment was to compare 2-meter vertical antennas made from materials used by amateurs, using methods to the best of our ability to obtain practical data. We recognize the professional literature and the way of doing business on model ranges. Presented here are our data to consider when making your choice between 1/4- and 5/8-wavelength vertical antennas.

*Not strictly correct mathematically but good enough for practical purposes. Editor

ham radio

antenna rotator

for medium-sized beams

Adapting TV rotators for postive directional control at reasonable cost

Most amateurs operating in the 10-15-20-meter bands use lightweight two- or three-element quads or three-element Yagis whose size and weight typically range from the Mosely TA-33 and CL-33 tribanders to the four-element 20-meter monobander such as Hy-gain's 204BA. These antennas weigh up to 40 pounds (18kg) and have a wind-loading surface area up to 7 square feet ($2m^2$). When looking for a rotator there's very little middle ground; these antennas are too large for a low-cost TV rotator but don't really justify the $100 to $200 cost of larger rotators (whose control systems leave something to be desired for operating convenience). For the mechanically inclined, the prop-pitch motor is an answer – if you can find one. However, these machines require mechanical modification plus design and construction of control circuits. This article describes an intermediate rotator system having superior operating features.

It isn't the weight of the beam, or even the rotator's ability to turn it, that's important in rotator selection; most TV rotators will *turn* much heavier beams than I've mentioned. Of more concern is braking resistance and lateral thrust, which involves the antenna's wind-loading area and turning radius. Wind forces and stopping torque determine what's needed for a rotator.[1]

Wind forces act in two directions on the rotator: a horizontal torque attempts to turn the beam off its heading (windmilling), and a vertical torque attempts to fold the rotator upon itself (lateral thrust). With enough space inside a tower, the lateral thrust can be transferred to the tower by placing the rotator inside it. But the forces attempting to turn the beam off its heading must be borne by the rotator. Note that it's not the steady, strong wind but gusts that do the most damage by causing the beam to oscillate. Following a series of articles published some time ago,[1] I installed a CL-33 tribander operated by a C-225 Alliance TV rotator. This system was at 40 feet (12m) surrounded by trees at 65 feet (19.8m), which provided some protection from wind. The rotator was backed up by an Alliance thrust bearing and was not inside a tower. Over a four-year period this rotator was just about equal to the task but wouldn't have long withstood an unprotected wind environment. Experience gained with this system encouraged me to find a way to improve it for an antenna system above the trees.

The design of the Alliance rotator is concentric with the mast, so a rotator system can be devised using two or more rotators. This arrangement, if the rotators are spaced a foot (30cm) or more apart, improves lateral thrust resistance while increasing the ability of the system to withstand windmilling. The additional rotators are mechanically in tandem, so the problem is to parallel their electrical control.

desired features

Some ideal features for a rotator system are:

1. An effective braking system.
2. Set-it-and-forget-it noiseless control.
3. Compass-rose control readout.
4. Automatic resynchronism of beam heading and control in case of rotator/mast slippage.
5. Avoidance of complex modifications or nonstandard parts.
6. Independent real-time position readout.

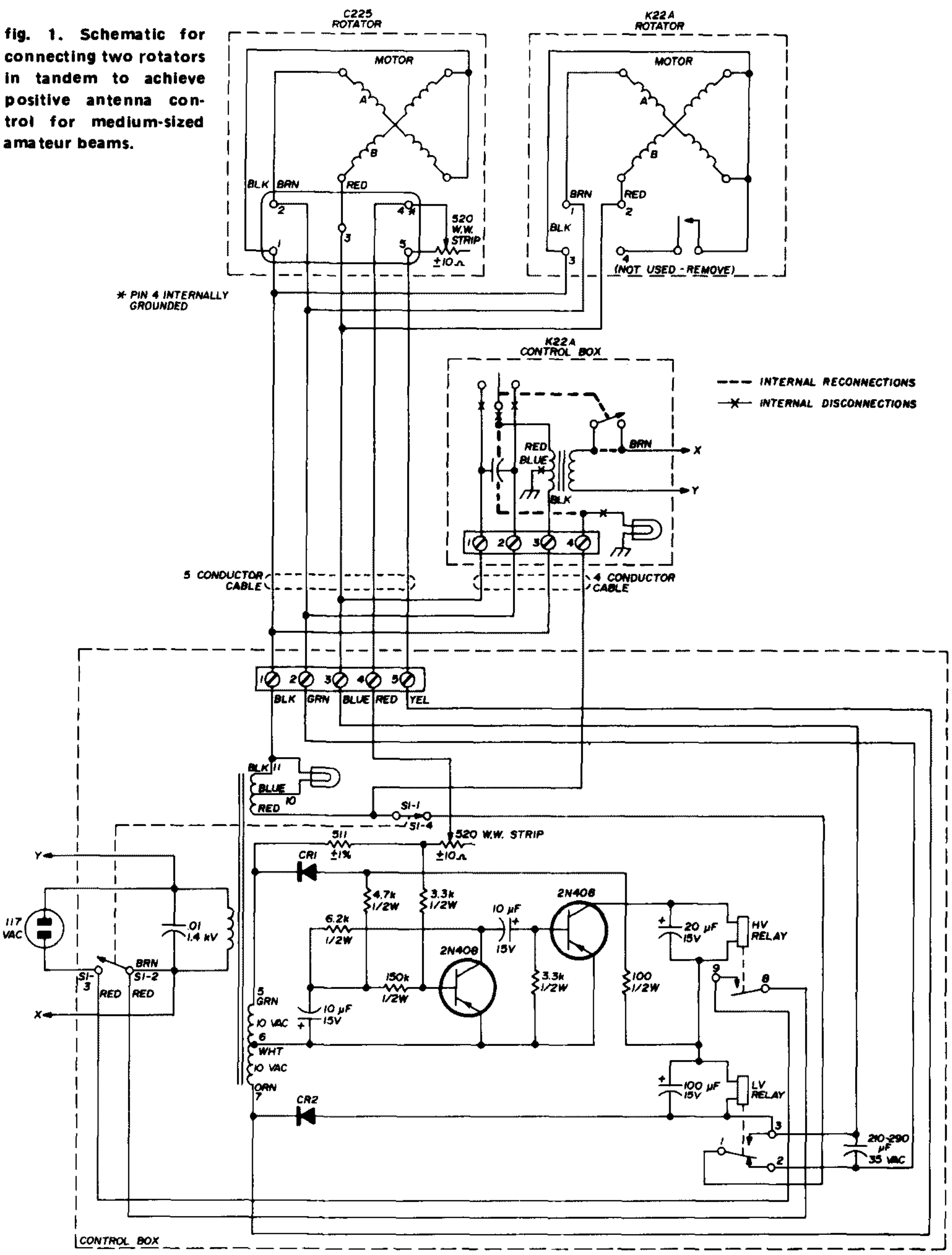
This is a tall order and isn't achieved in most amateur installations except very expensively. Using parts from Alliance TV rotators will fulfill the first five requirements; by bending the rule for special parts you can go all the way.

The mechanical-drive arrangement for the Alliance rotators discussed in this article is identical; their braking system is very effective. Not only does the gear train use a worm gear, effective in itself, but through a clever arrangement the motor at rest presents more resistance to rotation then when operating. I don't believe this rotator can be forced to turn without permanent damage. (Don't pin the rotator shaft to the mast!)

For wee-hour operating and DXing I prefer a control

By Forrest E. Gehrke, K2BT, 75 Crestview, Mountain Lakes, New Jersey 07046

fig. 1. Schematic for connecting two rotators in tandem to achieve positive antenna control for medium-sized amateur beams.



that doesn't make a lot of clacking noises, can be set to a new heading without keeping a finger on a switch until the beam gets there, and allows a quick visual determination of great-circle headings (dial in the form of a compass rose).

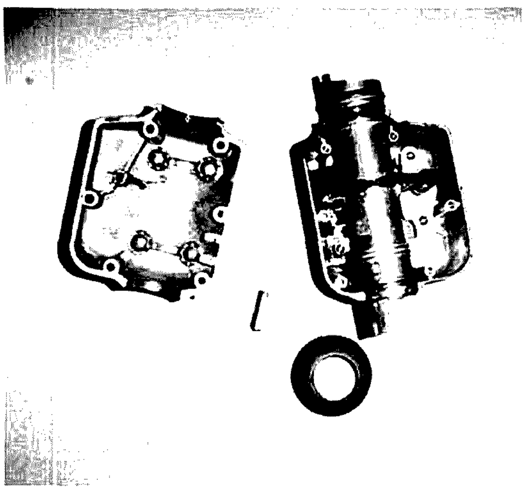
description

The Alliance model C-225* includes a control that meets these criteria. The control circuit is a transistorized bridge using in one of its arms a resistance element proportional to heading. As long as the rotator shaft maintains itself in the same position with the mast, synchronism of control and beam is automatic and reproducible. The circuit is highly reliable, in fact more so than a similar control for a much higher-cost rotator designed for larger arrays.

For additional rotators you don't have to make the same investment since only one control is needed. The Alliance model K-22A, also available from Lafayette,

*Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, New York 11791.

will provide an additional rotator and the required extra transformer and motor capacitor. These parts are connected into the C-225 control as in fig. 1. The K-22A control is a small plastic box and isn't objectional in appearance, so I left the needed components in their original positions and disabled the control switch by removing and taping up the leads. Only three connections must be made within the C-225 control. All other interconnections are made between the numbered control terminals. Be certain you color-match connections correctly between the two transformers and that you follow the numbered terminals in the schematic. This must be done to obtain correct supply phasing and proper motor rotation direction. It's a good idea to hook up this system on the bench with rotators not mechanically joined to ascertain proper operation. Avoid operating only one rotator with two motor capacitors in the circuit or the motor will rapidly overheat. The C225 rotator will determine your rotation stop direction, so be sure to return its rotation position to a known direction before mast assembly. (Most amateurs in North America use south as the stop direction.)



Disassembled Alliance K-22A rotator with T-stop removed.

Adding one additional rotator to the C-225 control didn't present any problems with its actuator switch or relays because of the additional current. However, increasing the number of rotators above two will require relieving the existing control switches and relays of this higher current. This can be done with additional relays or with triacs having higher current-handling ratings, but it becomes a bit more complex and the choice is left to the knowledgeable experimenter. Because the drive-mast diameter is only 1-3/8 inches (3.5cm), I don't recommend an antenna load needing more than three driving rotators. For example the inertia of a large array such as a Wilson 520 was difficult to manage with such a small drive mast, even though the rotators were inside a tower and adapted to a larger-diameter mast above the rotators.

The stop in the K-22A rotator must be removed because the relative position of the two rotators is not important. Carefully pull off the plastic collar on the top of the rotator, remove the six screws in the wells around the housing, and lift off the smaller half of the housing. Lift out the T-shaped stop and also remove the two contact springs from the terminal board. Reassemble, making sure that the original wiring color code is maintained. If the stop were not removed, each rotator would have to be in exactly the same rotational position at the time of assembly to the mast. Furthermore, if one rotator slips relative to the other, their changed stop positions would no longer allow full 360° rotation. Note that should the C-225 rotator slip (the one with the sensing potentiometer) the antenna heading will no longer agree with the control. If the change in rotation stop direction is only a few degrees, the fix for this is to pull off the control pointer and push it back so that the pointer agrees with the changed antenna heading.

After a high wind, an observation is recommended as you may find that the rotation-stop direction has slipped, or that rotator operation will result in wrapping too much coax around the mast. In this case there's no alternative to correcting the problem at the rotator. However, if your location is free of high winds or protected, this two-rotator system is quite adequate and economical. I operated such a system for two years, accepting an occasional small change in rotation stop direction by changing the control pointer. These changes tended to cancel out over a period of time. This inconvenience can be overcome making the system independent of rotator/mast slippage while using this slippage to advantage.

I've been discussing the problem of slippage and yet I've cautioned against pinning the rotator/mast coupling. Why not pin all mechanical couplings? Pin all except one, but leave one sufficiently weak link in the drive chain, preferably the rotator/mast coupling. This is insurance for all your effort and expense against unusually high winds. If something *has* to give, it should be here where no harm is done. Also, the oscillation induced by wind gusts and the sudden starts and stops of normal operation will produce early fatigue failure if the rotator clamps are too tight or have been pinned.

When unattended, it's good practice to leave the array pointing in the direction of least resistance to prevailing storm winds, making sure your rotation stop is located well away from that direction. Determining the position of least resistance is a matter of experimentation with your particular array. For the Mosley CL-33, it's with the ends of the elements pointing into the wind, but for some arrays you may find the boom must be so aligned.

automatic resynchronism

While attempting to build a system of four rotators to drive a stacked five-element array for 15 and 20 meters, I burned out the motor of one rotator on the bench by forgetting I had four capacitors across the single motor. This rotator housing was resurrected and used only for the sensing potentiometer and the mechanical rotation stop. The motor and gear train were removed, leaving

nothing to prevent this shaft from rotating with the mast within the 360° turning range. Now the driving rotators can slip relative to each other and to the mast in any amount, but as long as the single mechanical stop isn't reached, automatic resynchronism is achieved the moment the control is actuated thereafter. If high winds should force the mast off the control heading, the sensing potentiometer position, which freely turned with the mast, will be changed. As soon as it's actuated, the system will turn the mast in whatever direction is necessary to rebalance the bridge, re-establishing identity between mast and control heading. An extra K-22A rotator is the cost for this feature, but it's effective and readily available. I now have a rotator system that has control operating features unobtainable with most commercial rotators. This system has been in trouble-free use for three years and with no need to readjust the rotation stop.

These rotators have been designed for a standard 1-1/4 inch (3.2cm) pipe drive mast. Actual OD of this pipe is 1-3/8 inches (3.5cm). The rotator shafts have an ID that allows a little more than 1/16-inch (1.5mm) clearance. Use of shim material is advised to avoid flexure fatigue failure of the die-cast metal shafts because of the slight eccentricity this clearance creates. I cut some rectangular strips of 0.030-inch-thick (1mm) aluminum sheet 1 inch wide by 4 inches (2.5 by 10cm) long, wrapped them around the mast, and forced them between mast and rotator shafts at each end before installing the clamps. An excellent service manual is available from Alliance.* The manual is highly recommended as source information on operation of the control and also contains a useful trouble-shooting procedure in the event of a problem.

independent real-time readout

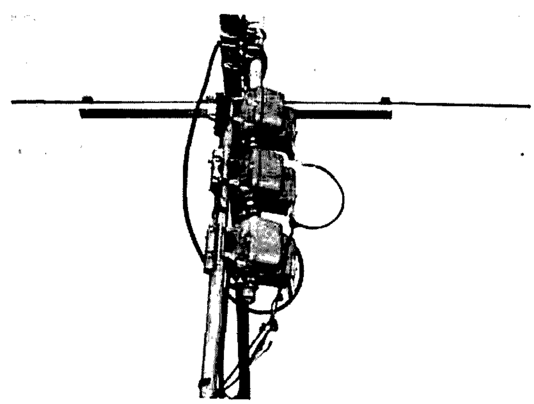
As with the previous additional feature, this refinement is optional. While automatic resynchronism is reliable, a readout independent of the rotator system is reassuring. Besides, it is sometimes useful to know the beam heading, even while rotating. Surplus 400-Hz synchronomotors were tried.* With a 7.5-volt 60-Hz supply a slight uneveness of motion (slot lock) was noted in the receiver synchro but was unobjectionable for the purpose. Because the transmitter synchro is driven directly from the mast, a special part is needed. The drive-wheel diameter must be identical to that of the mast. It was machined on a drill press from a piece of 1/4-inch (0.6cm) Lucite. My approach was to make the diameter 1/32 inch (1mm) undersize, machine a groove into the rim, fill the groove with silicone rubber, then smooth to size and cure. Silicone rubber is the only friction-drive material I've found that withstands weather. I used the same material to seal the cable opening in the motor, which is exposed to the weather. Note that the synchro motor is mounted on a metal bracket flexed so that it maintains drive-wheel pressure against the mast.

*Alliance Manufacturing Company, Inc., Alliance, Ohio 44601.

†Number SP-34 available from Meshna, Post Office Box 62, East Lynn, Massachusetts 01904.

The hookup provided with the synchros requires any one of the windings to be reversed when friction driven, so that the receiver synchro rotates in the same direction as the mast (the transmitter synchro turns oppositely). Of course, you could mount the transmitter synchro upside down, in which case the hookup should be followed as is.

A means of attaching the synchro through a small flexible shaft to the mast where it protrudes from the bottom rotator was considered but not tried. It is, of course, the better solution since it precludes slippage.



Two-rotator antenna drive with third rotator housing for potentiometer and rotation stop.

When an ice storm caused such slippage, I merely shifted the position of the receiver-synchro pointer to the new direction. I used the minute hand from a discarded clock for a pointer; however, any lightweight material may be used. The receiver synchro is mounted behind a great-circle map centered on New York City. At a glance I can see the antenna heading, whether moving or stationary. During high winds I can see the beam slipping position, a good indication that the antenna is not properly pointed into the wind.

conclusion

Being interested in electronic servo controls, I found this project to be fun. Out of it I formed a high respect for wind forces, especially after watching an 80-foot (24.4m) free-standing tower go down with my two five-element Yagis aboard (it's now securely guyed).

Most amateurs plunk down their money for a rotator, take what's commercially available, and accept the lack of convenience in control. The market doesn't warrant sophisticated control at a price most amateurs would be willing to pay. I've used low-cost automatic control aimed at a larger market and adapted it to popular amateur beams. Modifications are minor, require no hard-to-obtain parts, and are well within most amateur skills.

reference

1. E. Laird Campbell, W1CUT, "Antenna Rotators and Indicators," *QST*, April, May, 1967.

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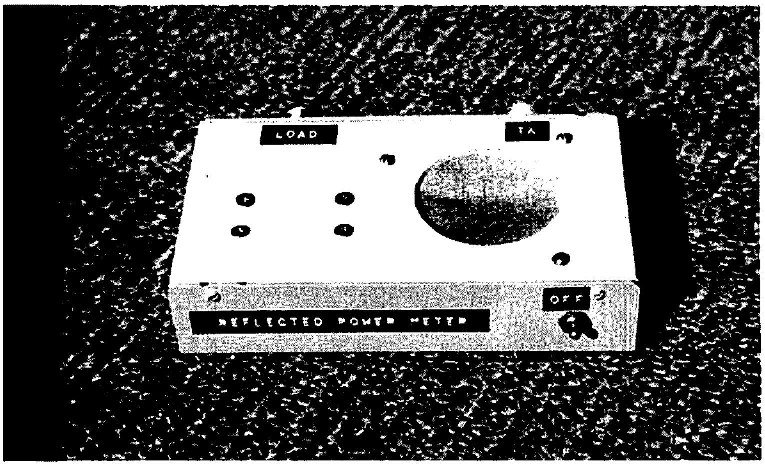


aural swr indicator

for the visually handicapped

Simple but effective impedance-matching circuit using an audio oscillator

A method of checking standing waves on a transmission line is a necessity in any amateur station. Most conventional circuits for measuring standing-wave ratio use the principle of the Wheatstone bridge or one of its variations and a high-impedance voltmeter. The voltmeter impedance must be high compared with the transmission-line impedance and also must have provisions for measuring the voltage applied to the bridge as well as the voltage applied to the bridge arms. Note that such a circuit is used to *measure* voltage standing wave ratio. Most amateurs are interested in obtaining the best possible impedance match between source and load, rather than measuring vswr, so much less sophisticated circuits than the swr bridge can be used.

A special problem exists for the blind amateur, who must rely on aural rather than visual clues to determine if the best possible match exists between source and load. The circuit shown here was built for a friend who had been using a noise bridge to adjust his matchbox antenna tuner. Not only has this swr indicator circuit saved him time, it has also helped him verify all tuning adjustments to his transmitter. This circuit is also useful for the seeing amateur, because swr can be checked aurally while watching meters.

circuit description

The swr indicator circuit is shown in **fig. 1**. An "inductive trough" transfers rf energy from the transmission line between source and load to a simple aural

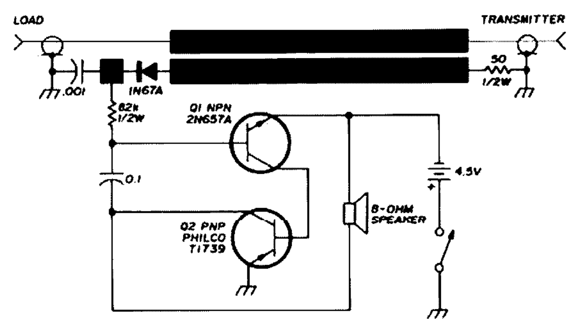
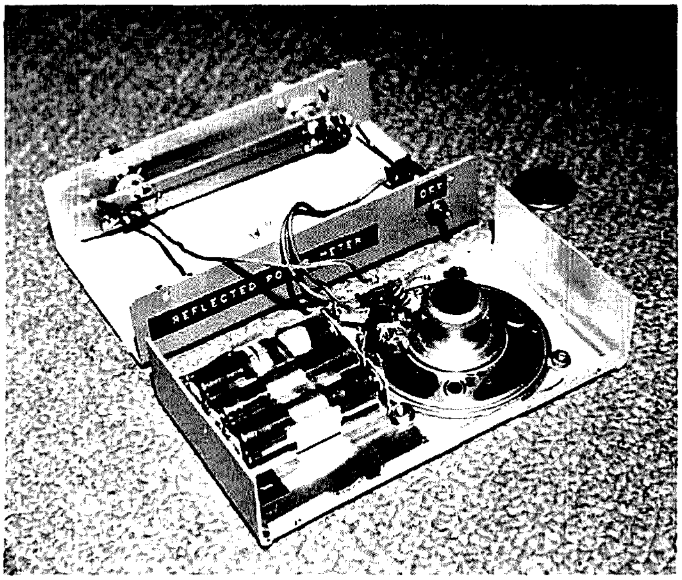


fig. 1. Aural swr indicator schematic. Darkened areas are foil strips that form an inductive circuit between source and load and audio circuit. Resistor in the base of Q1 may be increased to lower the audio tone

By Charles G. Bird, K6HTM, 875 Lindo Lane, Chico, California 95926



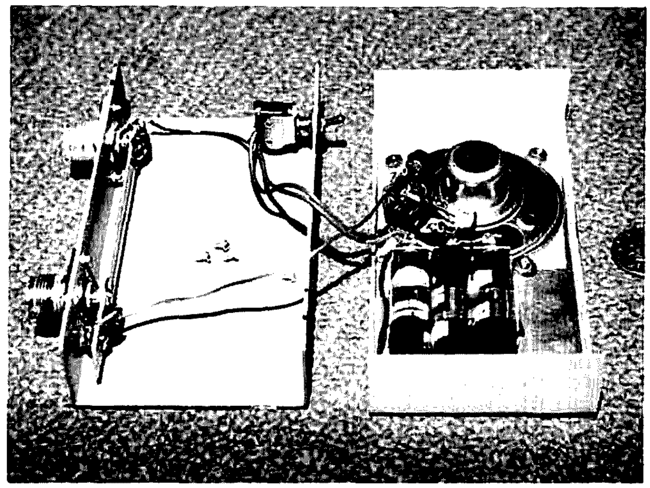
Inside the swr indicator. Foil strip inductive trough is in Minibox cover. Electronic components are wired directly to speaker terminals.

monitor.[1] The trough consists of two foil strips about ¼ inch (6mm) wide and 2-3/4 inches (70mm) long. They are spaced 1/16 inch (1.5mm) apart. One of the strips is the main conductor between coax connectors from source to load. The other strip transfers rf energy from the first strip to a germanium diode, which rectifies the rf energy to a dc voltage. This dc voltage changes the

W6HNL holds the swr indicator, which is connected between his transmitter and antenna tuning unit.

bias on the base of Q1, which causes the tone from the speaker to increase in pitch with increasing voltage.

The two transistors, 0.1 μF capacitor, and 82k resistor are mounted on the speaker terminals. Q1 and Q2 were salvaged from a surplus PC board. Their main requirements are reasonably high gain to establish oscillation and large enough dissipation to cause a loud tone. With the circuit constants shown, the idling tone is about 500 Hz. All components including three penlight batteries are enclosed in a 5½x3x1¼ inch (14x7.6x3cm) Minibox, LMB CR531.



operation

Ideally only the reflected wave is sampled so that minimum swr will exist when the transmitted tone matches the idling tone. If the transmitter and load cables are transposed, a forward reading will be heard as an increase in pitch. In practice, especially on the higher frequencies, it won't be possible to match the reflected tone with the idling tone because of stray capacitance. This is an advantage when using a matchbox since it's easy to peak the transmitter for maximum output on the rising pitch, then tune the matchbox for minimum swr on the descending pitch. The indicator may be left in the circuit at all times.

Thanks to W6HNL for his enthusiastic approval of the device.

reference

1. James Wiesmueller, WA9OHR, "A Simple CW Monitor," *ham radio*, January, 1971, page 65.

ham radio

vhf/uhf techniques

Joe Reisert, W1JAA

feeding and matching techniques for vhf and uhf antennas

If two or more amateurs get together in one place, more than likely antenna performance is one of the subjects to be discussed. This should come as no surprise because amateur antennas usually do double duty — working on both receive and transmit. Therefore, any performance degradation affects both the receiver and transmitted signal strength. Those amateurs who have good antenna systems not only receive best, they also radiate the strongest signal (all other things being equal).

A full discussion of antennas is obviously beyond the scope of this column, so I will concentrate on antenna feeding and matching techniques — both of these things affect not only the homebrew specialist, but also the owner of a commercial antenna. And, although this column is directed primarily to the vhf and uhf enthusiast, many of the same techniques are equally applicable on the lower frequency bands.

feed systems

The principal feed systems used on vhf/uhf antennas (see **fig. 1**) are the split dipole, folded dipole, delta match, tee match, gamma match and log periodic. Each of these has its own advantages and limitations, not to mention the personal preferences of some users; I'll discuss each of them and you can choose the one that you prefer. The split dipole (**fig. 1A**) is about the simplest way to feed a Yagi. Its major drawbacks are the need for insulating the feed from the boom and its low feedpoint impedance (typically 15 to 25 ohms) when this system is used to excite a Yagi beam. Hairpin loops or matching stubs are often used for impedance matching, but a balanced feedline or a balun is still desirable (more on this later).

The folded dipole (**figs. 1B** and **1C**) solves some of the problems of the simple split dipole and is a choice of many active vhf operators. One reason for its popularity is that the center of the folded dipole, opposite the feedpoint, can be directly connected to the boom. The arrangement shown in **fig. 1B** multiplies the nominal feed impedance by four times so it provides a more convenient impedance match to popularly available transmission lines.

If the input impedance of the simple folded dipole doesn't provide an impedance match, the variable ratio scheme of **fig. 1C** can be used. By changing the spacing, *S*, and/or the diameter ratio, $d1/d2$, this system can be used to provide a match to a variety of different transmission line impedances (see reference 1).

Above 225 MHz many amateurs have experienced problems with the folded-dipole matching systems shown in **figs. 1B** and **1C**. W1HDQ has discussed this problem in some detail[2] and offered the matching scheme in **fig. 1D** as a possible solution. Ordinary flat straps of metal can be used for element $d2$.

The folded dipole and its several variations can be used to match most antenna impedances. It's efficient, works well with popular balanced transmission-line impedances (200 to 300 ohms), and is easy to duplicate once a match has been obtained. The primary disadvantage of the folded dipole is the cut-and-try approach which is required when you have to match a new antenna.

The delta match down in **fig. 1E** is actually an extension of the W1HDQ's folded dipole in **fig. 1D**, is easy to build, and can provide a match to a wide range of balanced feedline impedances. The feedpoint impedance is adjusted by varying lengths $\ell 1$ and $\ell 2$ or the ratio $d1/d2$. However, some radiation can occur if $\ell 1$ and $\ell 2$ are large with respect to wavelength. This may lower antenna efficiency and increase side lobes, or both.

The tee match in **fig. 1F** is similar to the delta match but is less prone to radiate. Adjustments to $\ell 1$ and S and/or the ratio $d1/d2$ will provide a match to most feedline impedances. Capacitors C1 and C2 are necessary to tune out the inductive reactance of the feed system. In some cases the length of the dipole can be shortened slightly to eliminate the need for C1 and C2. The tee match is also designed for balanced feedlines and physical restrictions may limit its use above 450 MHz. The principal disadvantage of this system is the requirement for capacitors (if used) which may be sensitive to power level and weather conditions.

The gamma match in **fig. 1G** is a simplification of the tee match and is designed for unbalanced feedlines so

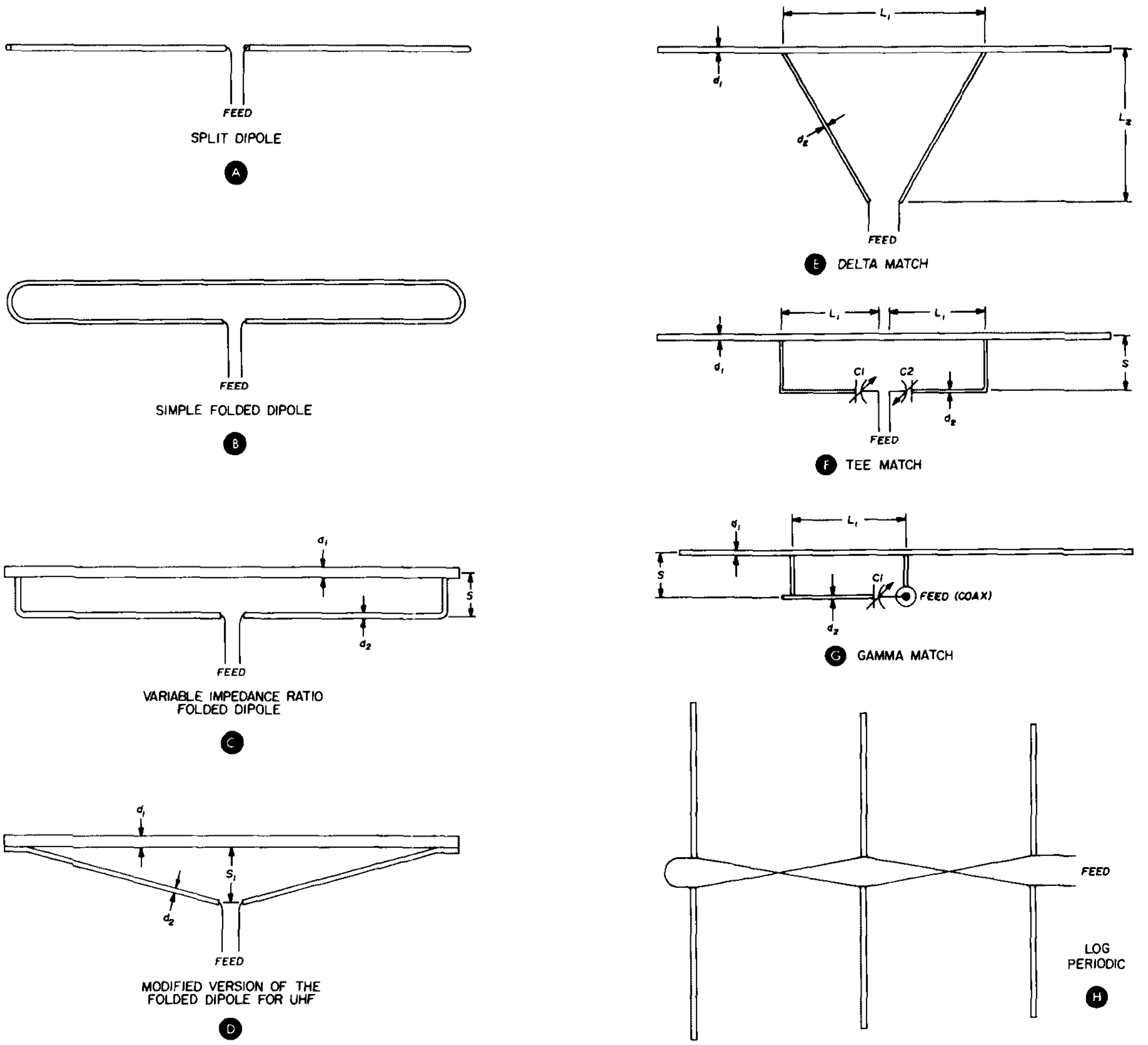


fig. 1. popular feed systems used on vhf and uhf antennas. The split dipole, folded dipole, delta match, tee match and log-periodic feed are balanced systems so a balun is required if coaxial transmission line is used. The gamma match (G) is an unbalanced system so can be used directly with coaxial cable.

coaxial transmission lines can be connected directly without a balun. The gamma match is also easy to use when different feedline impedances are desired (such as is required with circularly polarized feeds). As with the tee match, the major disadvantages of the gamma match are the capacitor and physical restrictions. A complete mathematical treatment of the gamma match is contained in reference 3.

The log-periodic feed in fig. **1H** is relatively new and is a spinoff from the log-periodic antenna.[4,5] Two of the advantages of this type of feed are the convenient feedpoint impedance (typically about 50 ohms) and wide bandwidth. Its disadvantage is the cut-and-try which is required when this system is used to match a new antenna. Also, it is a balanced feed.

feedlines

No vhf antenna article would be complete without a short discussion of feedlines, so I will digress for a moment and discuss some important feedline considerations. The main objective when choosing a transmission line is to keep the loss as low as possible, but unfortunately this creates a big tradeoff in cost vs performance. However, you should give due consideration to the benefits of a low-loss transmission line, especially if a very long feedline is required. Why put up a large, high-gain

antenna array and then lose half or more of your transmitter power (or received signal) because of high transmission-line loss?

Most amateurs are well aware of the fact that antenna height is a big factor in vhf/uhf performance. However, there is a law of diminishing returns. Once the antenna is

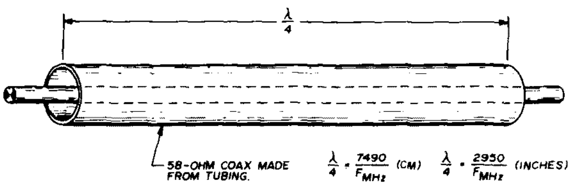


fig. 2. A quarter-wavelength coaxial transformer can be used to match a 50-ohm coaxial cable to semi-rigid 70-ohm CATV transmission line. The diameter is not important, but the ratio of the inside diameter of the outer tube to the outside diameter of the inner conductor should be 2.6:1 for a characteristic impedance of 58 ohms.

high enough that it clears most local objects and has a clear horizon, further increases in antenna height (and hence performance) may be offset by the extra feedline loss necessary to achieve that height. Therefore, put your antenna only as high as is absolutely necessary.

In recent years the trend has been toward coaxial cable instead of open-wire line or twinlead. This is because coax is easy to handle, doesn't radiate very much power, and is compatible with most modern vhf equipment (50-ohm transmitters, receivers and vswr bridges, etc.). In addition, more and better coaxial cables are becoming available, particularly on the surplus market.

So far as choosing a particular coaxial cable is concerned, small diameter RG-58/U should not be used, especially above 30 MHz. RG-8/U can be used in short runs but should be the foam-filled type with a good, full-coverage shield. Many of the bargain RG-8/U cables have only 95 per cent shielding (or less) so are very lossy at the higher frequencies. RG-8/U is particularly adaptable for running a line from an antenna, around a rotator, and to a tower. However, a lower loss transmission line should be used from the top of the tower to the radio shack.

The larger RG-17/U which is popular with many vhf/uhf operators can cause problems in areas with large temperature variations (such as New England). The expansion coefficients of the inner conductor and shield are different, so the center pin on the coaxial connector may pull back during cold weather, opening up the circuit to the antenna. This can be partially eliminated by rigidly mounting the end with several clamps (such as hose clamps). Another possible solution to this problem is the use of a type-LC coaxial connector in conjunction with an LC to type-N adapter. Type-LC connectors have more overlap on the center pin and are less likely to completely disengage in cold weather.

Semi-rigid, foam-filled coaxial cable in the ½-inch (13mm) and 7/8-inch (22mm) sizes is highly recommended. *Heliax,** especially the air-filled type, is probably the most desirable. Semi-rigid and *Heliax* cables are usually expensive but they exhibit very low loss and have a long life expectancy, so they pay for themselves both in performance as well as in the years of service they render.

One real bargain in coaxial cables has become available in recent years because of the increased number of CATV installations. These installations often use low-loss, semi-rigid, foam-filled 70- to 75-ohm coaxial cable (in both 1/2-inch [13mm] and 7/8-inch [22mm] sizes). In many cases the unused ends of reels of this cable are given away or sold at low prices by the CATV companies. The connectors for this cable, although expensive, are not as high priced as those sold for popular semi-rigid 50-ohm coax. If you're concerned about the use of 70-ohm transmission line, a simple quarter-wavelength coaxial transformer made from 58- to 60-ohm coax will easily transform a 70-ohm line down to 50-ohms (see **fig. 2**).

Twinlead is also an acceptable feedline for vhf use. However, only the low-loss outdoor types should be used. Federal K200, when available a few years ago, was very useful. The only major problem with twinlead is its response to moisture – even the slightest amount of dampness can adversely affect both vswr and line loss.

Although skipped over quickly, open-wire transmission line, if properly installed, is useful at vhf and uhf. The line insulators should be spaced about 5 to 7 inches (13 to 18cm) apart and should be made of a low-loss material such as 1/4-inch (6.5mm) Teflon rod. Line spacing should be small in proportion to wavelength for minimum radiation (0.75 inch or 19mm maximum at 432 MHz, proportionally smaller at higher frequencies).

The best places to use open-wire transmission line are in matching and feed harnesses in large arrays and where the line will not be subjected to pulling or twisting. To prevent radiation losses the vswr on the lines should be below 3:1. Open-wire lines in multiple lengths of one-half wavelength are excellent for coupling between baluns and matching stubs (this subject will be discussed in detail in a future column).

Before leaving the subject of feedlines it should be noted that low vswr at the antenna is highly desirable at vhf and uhf, especially when line losses are high (greater than 1 or 2 dB). A 3 dB feedline loss will increase to 4 dB if the antenna vswr is 3:1, and to 6 dB for a 7:1 antenna vswr! This condition may go partially undetected because a lossy line will dissipate part of the reflected power from the antenna so a vswr indicator located at the transmitter end of the line will not see the true vswr at the antenna. The solution, of course, is to have a well-matched antenna and low feedline loss.

baluns

The present trend, as noted above, is to coaxial transmission lines. On the other hand, most of the feed systems which are used at vhf and uhf are designed for balanced feedlines (**figs. 1A** through **1F** and **1H**). If an un-

Heliax* is a registered trademark of the Andrew Corporation.

balanced coaxial transmission line is connected directly to a balanced feed system there is a *possibility* that the transmission line will radiate energy, distorting the antenna's radiation pattern and possibly lowering gain. I use the word "possibility" because if a system is perfectly matched, there should be no need for a balun (per recent conversations with several antenna experts). Unfortunately, a perfect match is seldom achieved, and if it is, it usually occurs only at one frequency. Therefore, the use of a balun, a balanced-to-unbalanced line transformer, is recommended. A good balun will prevent rf currents from flowing on the outside shield of a coaxial line, thus forcing all rf current to be confined inside the cable. Although you may have heard differently, suffice it to say that the use of balun should not degrade antenna performance — it may actually improve performance.

Baluns come in various shapes and sizes as illustrated in **fig. 3**. One of the simplest types is the so-called "bazooka" or 1:1 balun shown in **fig. 3A**. In this balun the quarter-wave line acts as a trap to any rf current on the outside of the shield. For best performance the spacing between the lines should be at least two to three times the outside diameter of the feedline.

Another form of bazooka balun, often referred to as the beer-can balun,[6] is shown in **fig. 3B**. Varying the impedance of the feedline for the last quarter wavelength (inside the balun), provides an impedance step-up or step-down which can be used to conveniently match a resistive impedance other than the feedline impedance.

The simple 4:1 balun shown in **fig. 3C** requires an electrical half-wavelength coaxial cable, preferably low-loss type (such as foam-filled RG-8/U). Since the cable is an electrical half-wavelength long, it will be shortened by the velocity factor, v_p, of the line used. Velocity factors for most popular cables are published in the ARRL *Radio Amateur's Handbook*.[7] The impedance of this

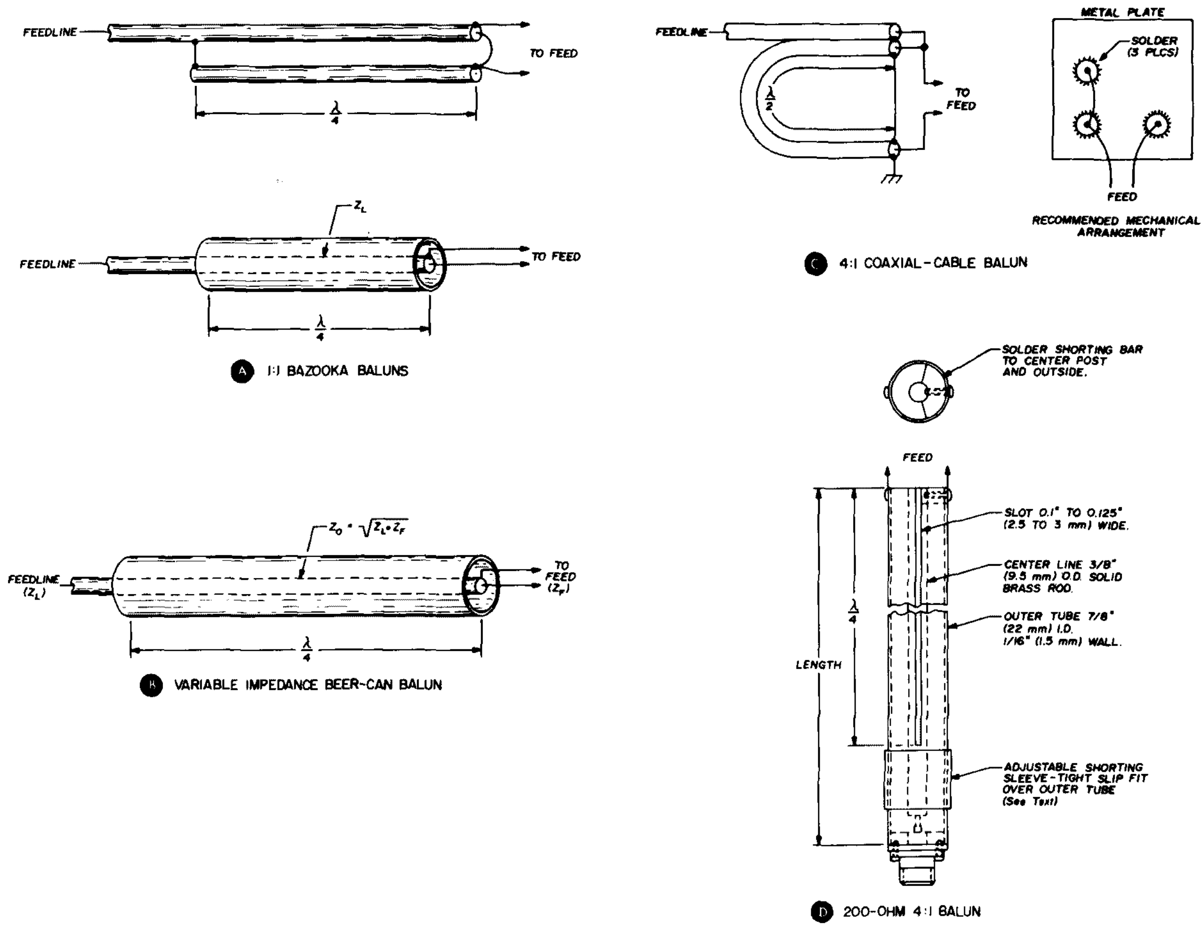
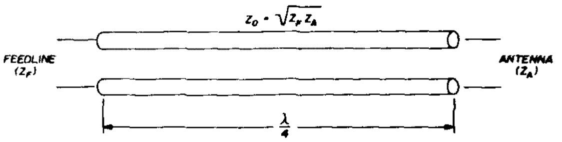
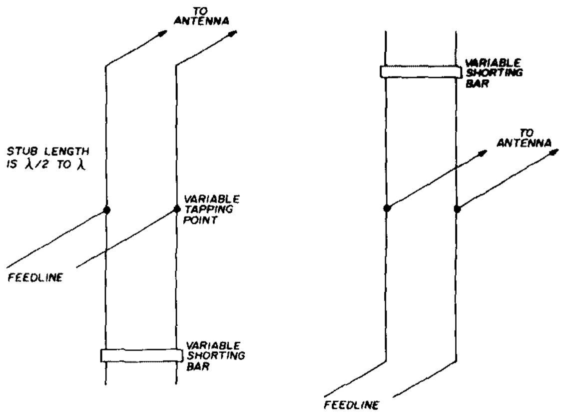


fig. 3. Various types of balanced-to-unbalanced transformers (baluns) which are suitable for vhf and uhf. Simplest type is the 1:1 bazooka balun shown in (A). The beer-can balun (B) can be used to provide an impedance step-up or step-down. The simple balun in (C) uses a half-wavelength of coaxial cable but the common ground should have low impedance; this can be accomplished with the metal plate shown to the right. The 4:1 balun in (D) can easily be set on frequency with the adjustable shorting sleeve and has no problems with undesired resonances.

line is not important for narrow-band work (less than 10 per cent bandwidth) common in vhf or uhf amateur operation. For wide bandwidth the impedance of the line should be equal to the geometric mean between the line and the feed impedance (100 ohms when using a



MATCHING OR Q SECTION



MATCHING STUBS

fig. 4. Matching stubs suitable for vhf and uhf. The Q section in (A) is often used if the antenna feedpoint is resistive (no reactance). The half-wave matching stub in (B) can be used to match virtually any impedance to another. The matching stub in (C) is used in those rare cases where the device to be matched is non-reactive and close to the impedance of the feedline.

50-ohm feedline and a 200-ohm antenna). The geometric mean is calculated with the following equation:

$$Z_g = \sqrt{Z1 \cdot Z2}$$

where Z_g is the geometric mean, and $Z1$ and $Z2$ are the line and feed impedances.

There is one very important consideration when using this simple balun: The ground connection at all junctions should have low impedance. In many cases the shields are simply twisted together and soldered – this causes loss and mismatches to exist. A better system is to make a small metal plate with three holes that will pass the outer braid of the cable as shown in **fig. 3C**. The braid is then dressed back and soldered to the metal plate. Use care to keep heat to a minimum by allowing the plate to cool off between soldering operations.

My favorite balun is the 4:1 type shown in **fig. 3D** which has been used by the National Bureau of Standards. This is essentially a low-loss 50-ohm coaxial line with two quarter-wavelength slots cut on opposite sides of the outer conductor. A simple shorting sleeve can be used to precisely set the balun on frequency. There are no problems with undesired resonances such as those sometimes associated with the beer-can balun (where more than one tuned element is used). In addition, the line in **fig. 3D** has no impedance steps so it is easy to build. For a more comprehensive list of balun types I would recommend reference 8.

matching networks

Since variable capacitors and inductors become a practical problem at vhf and uhf (capacitors become inductive, and vice versa), impedance matching at these frequencies is usually accomplished with tuned lines or matching stubs as shown in **fig. 4**. Matching stubs are both versatile and very efficient.

The familiar Q bar in **fig. 4A** is very handy if the antenna feedpoint is resistive (no reactance). The spacing of the rods and rod diameter are set by the geometric mean of this feedline impedance, Z_F, and the antenna feedpoint impedance, Z_A. A quarter-wavelength 283-ohm section, for example, would match a 200-ohm feedline to a 400-ohm antenna.

Usually the antenna or feed system (such as used on a collinear antenna) is not purely resistive so it's necessary to use a matching stub as shown in **fig. 4B**. A stub which is a half-wavelength long can match virtually any impedance to any other impedance with good vswr. In that rare case where the device to be matched is non-reactive and near the impedance of the feedline, the lines can be swapped as shown in **fig. 4C**.

matching techniques

With discussions of feedlines, baluns and matching networks out of the way, we can finally start talking about some matching techniques. The most popular

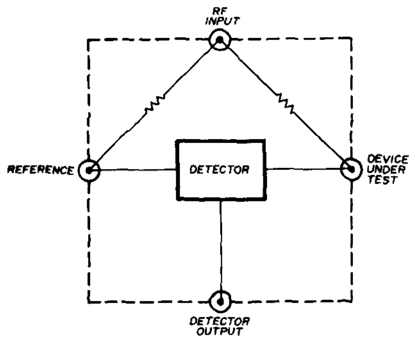


fig. 5. Simplified schematic of a vswr bridge for vhf and uhf. In use a reference load (usually 50 ohms) is compared to the device under test. If they are equal there is no output from the detector.

methods make use of a reflectometer, slotted line, network analyzer, vswr bridge or hybrid directional coupler. As with feed systems, each method has its own distinct advantages and limitations.

The reflectometer or moni-match is a form of directional coupler which uses either wire loops or a torodial

pickup. Although this device is inexpensive and works well below 50 MHz or so, at uhf it is frequently too crude and many require *watts* of power to operate.

The slotted line is one of the old stand-by vswr measurement techniques. By moving a probe along a partially

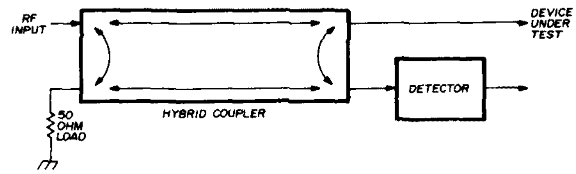


fig. 6. Hybrid coupler provides a convenient method for measuring impedance mismatches. When the device under test is tuned for minimum detected output a 1:1 vswr is virtually assured.

exposed transmission line the vswr can be interpreted by measuring the volt peaks and valleys along the line. Also, with careful calibration and some simple arithmetic any reactance presented by the antenna can be measured, then transferred to a Smith chart for interpolation. A slotted line is excellent for making precise, low-power vswr measurements. Its main drawbacks are size (at least one-half wavelength long) and use. For use on 432 MHz, for example, the slotted line must be at least 13 inches (33cm) long but a 26-inch (66cm) line is preferable. This limits the use of the slotted line to frequencies above about 400 MHz.

In addition to its size and weight, the slotted line is cumbersome to use because you must move the probe carriage back *and* forth after each matching change to see if you have made an improvement. Slotted lines that are suitable for amateur measurements are described in references 9 and 10.

The network analyzer is a powerful test instrument which has become very popular among microwave engineers during the past ten years. It directly displays vswr on a Smith chart and is easy to use. The primary drawback is cost, typically $10,000 or more, so its use is limited to those who have access to it in a professional laboratory.

The vswr bridge has been around for a long time, but it hasn't attained the popularity which it deserves. It's inexpensive, simple to build, and easy to use. Several are available on the commercial market. A simplified schematic is shown in fig. 5. This device works on the same principle as the low-frequency bridges which are common to test equipment. If an rf input is present, the power will divide between the reference and the device under test. If the reference (usually a non-reactive 50-ohm load) and the device under test are equal, the detector output will theoretically be zero. If the device under test is not equal to the reference, a voltage will be present at the detector output.

To use the vswr bridge to match a 50-ohm antenna, all that is necessary is a non-reactive 50-ohm reference load. You simply tune the antenna for minimum voltage *from the detector. The beauty of the* vswr bridge is that it can be used to measure various impedance levels simply by changing the reference load. Also, by substituting a known mismatch load for the device under test, the vswr can be quickly determined by interpolation.

My favorite uhf matching technique is based on the use of a hybrid directional coupler (**fig. 6**). If the antenna is properly matched, theoretically no power will reach the detector. However, even the slightest vswr will cause a detected output. All you have to do is tune your antenna for minimum detected output and a 1:1 vswr is virtually assured.

In this column I have discussed feed systems, feedlines, baluns, matching networks and matching techniques. Low-loss feedlines and low vswr have been stressed for obvious reasons. In a future column I will discuss measurements more thoroughly, and will recommend some easy-to-build test equipment which you can use to evaluate your own antenna system. Hopefully this article will inspire you to do more work on your antenna system and hence improve the performance of your vhf/uhf station.

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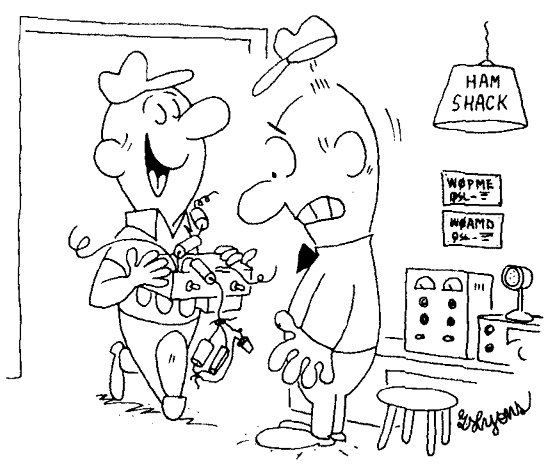

"Still think you can fix anything?"

comments

antenna gain

Dear HR:

Antenna gains still seem to be an area of fuzziness. In microwave work it is proper and useful to refer power gain to an isotropic radiator in free space.[1] However, at high-frequencies, which concerns the majority of amateurs, confusion sets in. The IEEE standard states:

For horizontal polarization . . . the reference antenna may be a half-wave dipole at the same height. In this case the gain of the reference antenna is taken as 8 dB, which represents the approximate sum of the 2.15 dB free-space gain of a lossless half-wave dipole and the 6 dB augmentation due to the assumed perfectly conducting earth.

In other words, a simple half-wavelength dipole, which we all know well and can put up for very little money, has a power gain of 8 dBi per IEEE Standard 149.

In some cases gain is referred to an isotropic radiator at the same height and foreground. In these cases (and you are not always clearly informed as to whether the free-space condition is being used as a reference), you should deduct 2.15 dB. However, if the reference isotropic radiator is in free space, 8 dB must be deducted.

For high-frequency amateur commercial work there is no reason to go to the free-space isotropic radiator and back again to the reality of the half-wave dipole. The practical basis of comparison is always a half-wave dipole at the same height and foreground. The earth is always there. Unless, of course, you want to make the antenna you are working on seem better than it actually is. To state gain relative to a half-wave dipole, deduct 8 dB from the gain figures given per IEEE Standard 149 measurements or calculations.[2]

Even by discounting 8 dB, you cannot be sure that you have brought dBi back to a reference which you clearly and unambiguously understand. The claimant may be talking about *directive gain*. This reference, which is permissable and is defined by the IEEE, is derived from mathematical integration of the antenna pattern, and you may find that the antenna patterns are calculated, *not* measured or proven. That is, if the claims are based on anything substantial. In some cases they may be deduced from a partial measurement of the full spatial pattern.

Under the directive-gain definition there can be large losses in the antenna and ground such that, for transmitting purposes, an 11 dBi log-periodic beam could lay down a smaller field than a garden variety, half-wavelength dipole.

The Electronic Industries Association has also introduced an antenna gain standard. EIA Standard RS-409[3] says:

The gain of a lossless half-wave dipole shall be used as a standard gain unit . . . and . . . the power gain of an antenna shall be expressed in dB over the gain of a lossless half-wave dipole, or dBd.

It would be helpful if the amateur antenna manufacturers, magazine publishers and advocates of a particular high-frequency antenna design would stick to the EIA dBd.

Now, concerning log-periodics and the gains mentioned in discussion of them. As an example, I rate the *ARRL Antenna Book's* presentation,[4] taken from the excellent work of K4EWG, as first rate. However, when the average amateur reads that a well-designed log-periodic dipole array has "approximately 7.4 dB gain over a half-wave dipole," he will skip over the discussion of the design constant, T, and the relative spacing constant, σ. He will also read that, "Tilting the elements toward the apex will increase the gain 3 to 5 dB." Wow, he thinks, 7.4 + 5 = 12.4 dB gain!

However, the truth is that he would be better off with two or three 3-element Yagis (one for each band). The National Bureau of Standards made measurements on a 26-element, double-curtain, high-frequency log-periodic[5] and found that beamwidths were about 70 degrees. Gains measured 4.5 to 6.4 dBd. Another report on log-periodics,

1. "IEEE Test Procedure for Antennas," Standard 149, *IEEE Professional Group, Antennas and Propagation (PGAP)*, May, 1965. Also see definitions of terms, *PGAP*, May, 1969, and IEEE Standard 145-1973, *PGAP*, January, 1974.
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from ITSA, is referenced to IEEE power gain.[1,6] Over the frequency range for 4 to 30 MHz the beamwidths were 39 to 70 degrees for the high-frequency log-periodic. Gains, converted to dBd by subtracting 8 dB, were 4.2 to 6.6 dBd.

Beamwidths for the VLP described in the ITSA report were 45 to 120 degrees over the frequency range from 6 to 30 MHz. Applying the 8 dB correction factor, the gains were −8.6 dBd to +2.9 dBd. The particularly disappointing results for the VLP are attributed to ground losses, end loading and unbalance, although the site (a rice paddy) and ground mat were probably much better than most amateurs could provide.

This is not to say that log-periodics are always bad. They pay their way, for example, as wideband television receiving antennas, or where full frequency agility is required as in some military applications. However, I don't think they can be justified for competitive DXing by amateurs.

Paul D. Rockwell, W3AFM
Chevy Chase, Maryland

non-synchronous impedance transformer

Dear HR:

The application of the non-synchronous impedance transformer in *ham notebook*[7] was especially interesting, and in my opinion, worth a further look. Unfortunately, the author did not discuss bandwidth except in very general terms.

Using the 2:1 transformation described by Mr. Keen, I checked his transformer against the conventional quarter wave for bandwidth at the 10% and 30% points as represented by lengthening (or shortening) the 28.13 degree length required for each section. The vswr as seen by the generator for the non-synchronous transformer is 1.14 and 1.52 against the quarter wave being 1.12 and 1.38, respectively, at 10% and 30% above the design frequency.

Though the bandwidth in terms of vswr is compromised, the difference is only about 2½ per cent when applied to the "widest" band, 75 to 80 meters. Perhaps the most profound advantage is that it requires one-third less space for matching when compared with the traditional quarter-wave section.

W5TRS is to be congratulated for bringing forth the good idea!

Raymond P. Aylor, Jr., W3DVO
Garrett Park, Maryland

7. Henry Keen, W5TRS, "Non-Synchronous Impedance Transformer," *ham radio*, September, 1975, page 66.

432-MHz Yagi

Dear HR:

The article on the high-gain Yagi for 432-MHz in the January, 1976, issue of *ham radio* indicates that some amateurs have not been successful in reproducing the WØEYE Yagi design. I would like to point out that in all the cases I've seen where a builder did not obtain the claimed performance, modifications had been made to the original design. *Precise duplication* is the only way that performance can be guaranteed. Changing the boom or element diameter, or especially the method of mounting the elements, can have disastrous effects on gain. Tests at the East Coast VHF Society in Trenton, New Jersey, in August, 1974, showed the WØEYE 432-MHz Yagi to have approximately 0.5 dB more gain than the K2RIW design. This is to be expected in view of the fact that the boom for the WØEYE Yagi is 0.7 wavelength longer than the K2RIW Yagi (4.2λ vs 3.5λ).[8]

Don Hilliard, WØPW (exWØEYE)
Boulder, Colorado

8. The original WØEYE Yagi design information published in the January, 1972, issue of *QST* contained some errors which were corrected in the *World Above 50 MHz* column in the March, 1972, issue.

microprocessors

Dear HR:

The microprocessor article in the December issue of *ham radio*[9] left me baffled. It compared "computers" with microprocessors, and it compared "programmable calculators" with microprocessors, but they aren't the same thing. A microprocessor with *memory and peripherals* can be either a "computer" or a "programmable calculator." A microprocessor-based system can handle the same sort of sophisticated mathematical computations that the authors imply is inherent in computers and programmable calculators, and in the same way – with a digital algorithm controlled by some sort of memory.

Ordinarily, because of the small word size and limited instruction set available in microprocessors, they are slower than most computers but they're faster than most programmable calculators. The data rate of five-hundred 16-bit samples per second is equivalent to one sample each 2000 microseconds. The modern microprocessors (8080A, 6800, 6501) run with clocks operating from 500 kHz to 3 MHz. To store a data value from a peripheral device, using the 8080, should require about 60 clock cycles. The faster 8080A runs with a 1-MHz clock so the time per sample is about 60 microseconds or 16000, 16-bit samples per second.

With Direct Memory Access (DMA), a goody which is nearly mandatory on *any* high-speed data acquisition system, the rate could be as high as one sample for each six clock cycles or 160,000 samples per second. This is faster than you can stuff the data into most mass-storage devices (tape, disk).

Microprocessors have advantages (low cost) and disadvantages (limited software) that leave the user with mixed emotions, but there is no clear-cut, qualitative distinction between microprocessor-based digital systems and computers or programmable calculators. Because of their low cost, you can think in terms of distributed systems with microprocessors ROM-programmed to perform specific functions (math, logic, control) and pipeline data to them. On the other hand, due to the paucity of useful software, you must realize that the programming costs will vary somewhere between enormous and staggering.

Microprocessor and memory costs are going down, microprocessor capabilities are going up, and software is very rapidly becoming more available and more efficient. *Today* microprocessor-based systems can give you one helluva bang per buck – a quantum jump in capability is in the wings.

C. E. Deckard, WB4FAR
Huntsville, Alabama

9. D. Larsen, WB4HYJ, Peter Rony, and Jonathan Titus, "An Introduction to Microprocessors," *ham radio*, December, 1975, page 32.

the ham notebook

low-cost two-meter colinear uses PVC pipe mast

The colinear described here is about as low cost as you can imagine for fixed-station use, costing less than ten dollars. Omnidirectional, the antenna consists of three half-wavelength stubs which can be fed directly with balanced 300-ohm line (or coaxial cable and a 4:1 balun).*

The colinear uses a 10-foot (3m) section of 3/4- or 1-inch (20-25mm) PVC water pipe as a mast and elements made from aluminum clothesline wire. PVC water pipe is available through discount and hardware stores and the Sears mail order catalog. CPVC pipe is also available but is much higher priced. A full line of fittings and special cement are available for capping and splicing. The local cost of a 10-foot (3m) length of 1-inch (25mm) PVC pipe is less than $2.00. While a three-element two-meter colinear will fit nicely on a 10 foot (3m) mast, I designed the antenna so about one quarter wavelength of the top element extends above the mast. This allows for additional room at the bottom of the mast so nothing shorts out.

To build the antenna, first unroll and straighten the aluminum wire and cut two 97.5 inch (2.48m) lengths. On one of the lengths of wire measure out 57 inches (1.45m) from the end and make a 90° bend. Make another 90° bend 1 inch (2.5cm) from the first. You now have a big hairpin with one leg 57 inches (1.45m) long and the other 39.5 inches (1m) long.

Lay the hairpins aside and drill eight 1/8-inch (3mm) holes in the PVC mast as shown in **fig. 1.** Mark both legs of the hairpin 19 inches (48cm) from the end. Now insert the long leg of the hairpin into the second hole from the top of the mast on the side opposite from the feedpoint. Put the shorter leg into the next hole and push both legs through until the marks appear. Now bend both legs of the hairpin back against the mast with the longer leg toward the top and shorter leg toward the feedpoint. Push the hairpin further through the PVC mast, bend the longer leg, and feed it through the top hole in the mast and out through the open end of the pipe. Now push the vertical elements back against the mast and secure them with vinyl tape. Form a small eye on the end of the shorter element and attach it at the feedpoint with a self-tapping screw.

The other 97.5-inch (2.48m) length of aluminum wire is used to build the lower half of the antenna. Use the same procedure as before except in this case the longer leg of the hairpin goes toward the bottom and both ends are secured with self-tapping screws. A center-drilled PVC cap slipped over the top of the PVC mast will keep undesired moisture out of the antenna. The quarter-wavelength stubs may be allowed to stick out at right angles, or they may be bent into a circle to ease handling.

On my own antenna I cut a large hole in the PVC mast at the feedpoint, split a PVC coupling lengthwise, secured the feedline and coaxial balun to the inside of the split coupling with machine screws, pushed the coax out the bottom of the mast, stuffed the balun inside, and cemented the split coupling over the hole. The ends of the elements were secured with machine screws and nuts. If you use this method, be sure to try the antenna out before you cement anything in place because the bond is *permanent.*

Although no extensive tests have been performed on this antenna, it appears to be equal to a commercial half-wavelength vertical. Best of all, it's inexpensive and requires no tuning.

Don Norman, K8LLZ

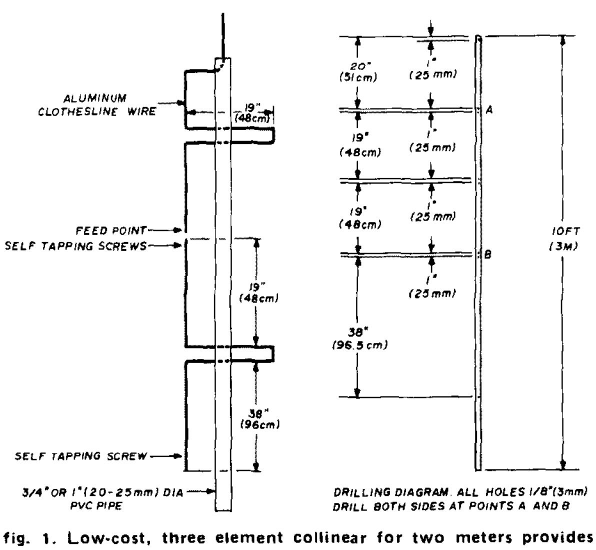


fig. 1. Low-cost, three element collinear for two meters provides omnidirectional coverage and can be built for less than ten dollars.

*Ed Tilton, W1HDQ, *The Radio Amateur's VHF Manual*, ARRL, Newington, Connecticut, 1972, page 156.

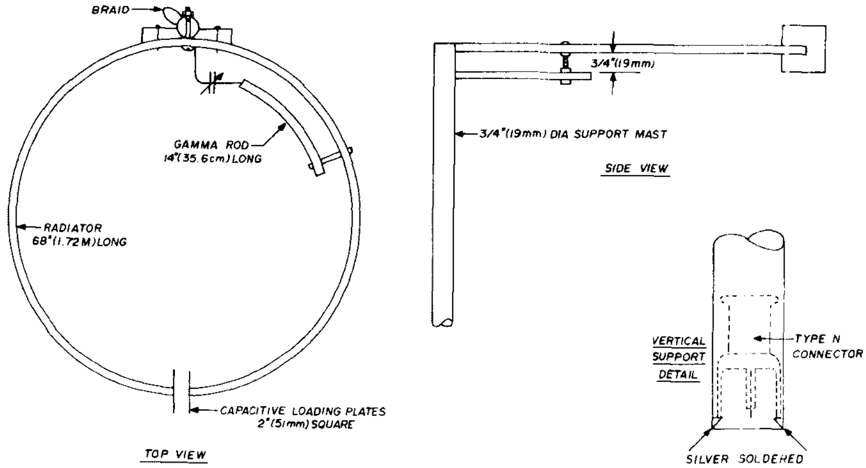


fig. 2. Full-wave halo antenna is mounted one-quarter wavelength above truck cab (or car roof). Radiator and gamma rod are 1/2" (13mm) aluminum tubing; support mast is 3/4" (19mm) copper water pipe.

mobile oscar antenna

Most amateurs aren't crazy enough to work a satellite while screaming down the freeway at 60+ mph, but if you have that desire let me suggest an antenna which is quite efficient for that type of operation.

The antenna described here was built primarily to achieve horizontal polarization for two-meter sideband operation. As a bonus I have found it works very well on the Oscar 6 and 7 mode-A uplinks. The conventional two-meter halo leaves a lot to be desired as an efficient omnidirectional horizontal radiator. I decided to use a one-wavelength loop and capacitively load the ends (see fig. 2).

For rigidity and simplicity I elected to use the gamma match. In order to achieve some gain I mounted the radiator approximately one-quarter wavelength above the roof of my truck. A type-N coaxial feedthrough was placed in the middle of the cab top. The quarter-wave supporting mast is made of 3/4-inch (19mm) ID copper pipe. A piece of RG-8/U coaxial cable with type-N connectors is placed inside the copper pipe, inset approximately 1/8 inch (3mm), and silver soldered in place.

The radiating portion is built from 1/2-inch (13mm) OD aluminum tubing 68 inches (1.72m) long and bent into a circle so the ends touch. Plugs, 3/8 inch (9.5mm) diameter and 1 inch (25mm) long are swaged into the ends and drilled to accept number 6 (about 3.5mm) sheet metal screws. The capacitive plates for the ends are 2-inch (51mm) aluminum squares or 2½-inch (6.5cm) aluminum disks. After attaching the capacitive plates to the ends of the radiator the spacing of 3/8 inch (9.5mm) is held constant by the use of ceramic insulators. The radiator may now be attached to the copper vertical support by using an element-to-boom insulator (25¢ each from KLM).

The gamma rod is 14 inches (35.6cm) long , has the same radius as the radiator, and is mounted below the radiator using two 3/4-inch (19mm) ceramic insulators. A 15 pF variable capacitor is mounted on the end of the gamma arm closest to the vertical support. Attach the center conductor of the coax that is coming out the top of the vertical support to the capacitor to place it in series with the gamma arm. The braided shielding may be attached to the vertical support on the center of the radiator.

Drill a hole through the gamma arm 2 inches (51mm) in from the end opposite to that where the capacitor is attached. Continue this hole through the radiator. Place a 1½ inch (3.8cm) long screw through these holes to short the radiator to the gamma arm at this point (a sliding short may also be used, possibly made from a pair of *Adel* clamps or small hose clamps and a small strip of flashing copper). This shorting point was found for 145.0 MHz ±1 MHz.

Tune up was done using a military TS-47 vhf signal generator, an HP-415 vswr indicator and a homebrew bridge detector head. The antenna is very sensitive to hand capacitance so a long non-metallic tuning tool should be used. A good quality vhf vswr indicator can also be used for tuneup with good results.

Doug A. Clingerman, W6OAL

portable magnet-mount antenna

The compact gutter-mount antenna, long popular with two-meter fm enthusiasts who have to travel, cannot be used on many of the newer cars because they're not equipped with rain gutters. Following is a description of an economical magnet-mount antenna (fig. 3)

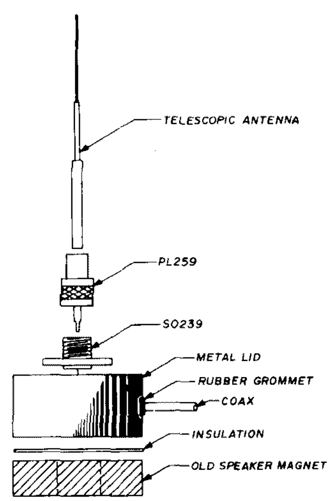


fig. 3. Simple magnet-mount mobile antenna can be built from easy to find parts, and can be tuned to 145, 220, or 440 MHz by adjusting the length of the telescopic antenna for minimum vswr.

which can be built in less than an hour from easy-to-find parts.

First solder a short length of number 12 (2mm) copper wire to the base of a short telescopic antenna, solder the other end of the wire to the center pin of a PL259 coaxial connector, then fix the telescopic antenna in place with 5-minute epoxy. Mount a SO239 coaxial socket on top of an old spray-can lid (be sure to include a ground lug for the coaxial shield connection). Drill a hole on the side of the lid for a 3/8 inch (9.5mm) rubber grommet and install a 6 foot (1.8m) length of RG-58/U coaxial transmission line. Now place some insulation inside the lid and shim the magnet so it fits tightly. I used glass tape which was held in place with fast-curing epoxy cement. The length of the telescopic antenna can be adjusted for operation on 145, 220,or 440 MHz, and can also be set for minimum standing-wave ratio.

Fred Snow, WB2YYU

7-MHz attic antenna

The 3:1 or 4:1 swr KH6HDM mentioned in his article on dipole antennas* would not pass the New York City Board of Education specification which requires an swr of 1.5:1 or better for radio and television receiving systems. This year I had 20-meter dipole in my attic which had a 1.25 swr. Since I wanted to operate on 40 meters, and this antenna is one-quarter wavelength long at 7 MHz, I short circuited the balun terminals and end fed the 32 foot (9.75m) length of wire – not unexpectedly, the swr was extremely high.

I installed an old broadcast-type 365 pF variable capacitor in series with the feedpoint but the vswr was approximately 2.8:1 at best. However, when the 365 pF variable was installed in place of the jumpered balun and adjusted for minimum reflected power the swr measured about 1:1. When a 4 foot (1.2m) length of 3/8 inch (9.5mm) copper pipe was used to replace wire which had bends in it, the swr dropped to nearly 1:1. A diagram of the complete installation is shown in fig. 4.

Allen Porterfield, W2ISL

*Albert Lee, KH6HDM, "Dipole Antennas," *ham radio*, November, 1975, page 60.

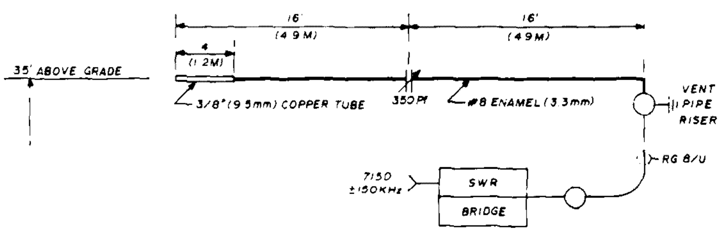


fig. 4. 7-MHz attic antenna is built from a 20-meter dipole. Swr is less than 2:1 over the entire 40-meter band. The 4-foot (1.2m) section of copper pipe may not be required if there are no bends in the antenna wire.

improved low-band inverted-vee installation

Many amateurs who own towers use some form of inverted-vee antenna on 80 and 40 meters. At my station I use an inverted-vee which is trapped for operation on 40, 75, and 80 meters.* When I first put the antenna up, however, the swr was very high on all bands and I got badly tromped in every DX pileup. A hip-pocket analysis suggested that the apex of the vee was too close to the tower – the center insulator was suspended by a two-foot (60cm) length of nylon rope about 6 inches (15cm) out from the face of the tower.

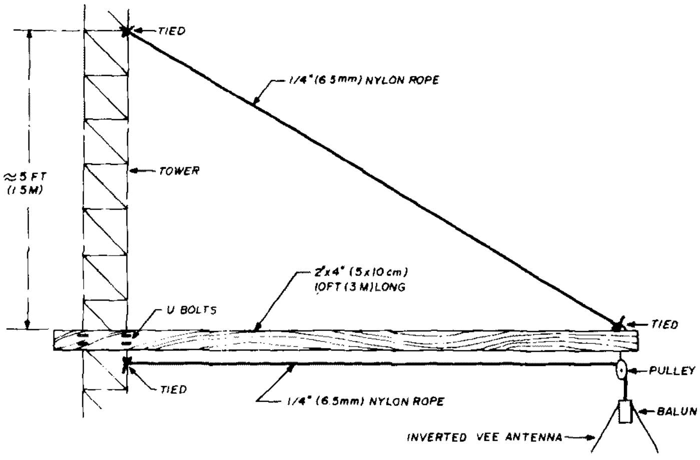


fig. 5. When inverted-vee antenna is mounted well away from the tower, as shown here, performance improves markedly.

To move the vee further out from the tower, I bought a 10-foot (3m) length of 2x4-inch (5x10cm) lumber which I bolted to the legs of the tower with four U-bolts as shown in fig. 5, and supported the far end with a length of 1/4-inch (6.5mm) nylon rope tied around one of the tower legs 5 feet (1.5m) above the wooden beam. A pulley was attached to the underside of the wooden beam at the far end so the inverted-vee could be lowered for repair and maintenance.

The result was very gratifying. Swr on all bands was reduced considerably, although further trimming was required for optimum results. While I continue to get tromped in the pileups by those members of the DX gang who use phased arrays or full-size ground planes, this new arrangement is far superior to dipoles and vees which are mounted close to the tower.

Bob Locher, W9KNI

improving the swr meter

The swr meter is one of the most used pieces of test gear. Many variations of the circuit have been devised and most suffer from inaccuracy. A

*Bob Polansky, W6JKR, "Low-Band Converted-Vee Antenna," *ham radio*, December, 1969, page 18.

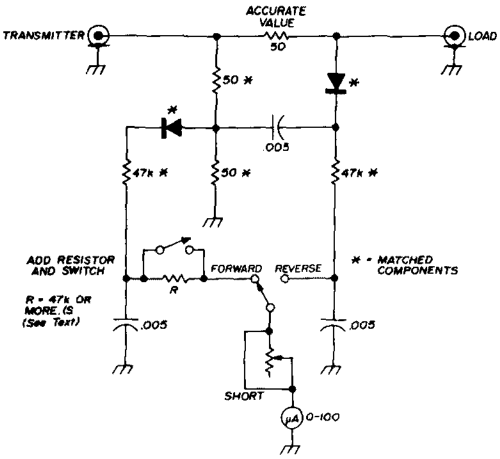


fig. 6. Swr bridge as shown in ARRL handbooks modified to provide greater accuracy at low readings.

considerable improvement can be made rather easily on the lower ranges of most meters.

Before you say, "Mine is ok," try this: Assuming a 50-ohm meter, place a 100- or 25-ohm resistor on the output and see what you get. Two 50-ohm loads on a T fitting will do. You should get a 2:1 swr reading or a reverse reading one-third of full scale. The relationship is full scale plus reverse, divided by full scale minus reverse. With a 100-microamp meter it is 100 plus 33.3 over 100 minus 33.3, which equals 1.99 or 2:1. If you get this, smile and try measuring with a different power level. If you're frowning now, here is what you can do.

Assuming good balance in the construction of the meter and adequate matching of the diodes and resistors, the major problem is in nonlinear resistance of the detector diodes. When measuring the lower swr conditions, much more current is used to calibrate the instrument in the forward position than is measured in reverse. Large series resistances and sensitive meters tend to even out this nonlinearity. However, accuracy suffers as readings get smaller and low readings will give an unduly optimistic impression of the match. Measurements at different power levels will also give different readings.

The bridge shown in **fig. 6** is from various ARRL handbooks and is meant for measurement at low power, not as an in-line device. As shown, a resistor is placed in series with the lead to the *forward* switch contact. A switch is installed across this resistor to restore normal operation. For these measurements a *short is* placed across the variable resistor or it is left in the most sensitive position. Transmitter power is adjusted for full-scale readings and should be limited to a few watts.

For a starting point use a resistor of about equal value to the one already in series with the *forward* diode. Install a load of 25 or 100 ohms on the bridge and make a measurement. Calibrate to full scale in *forward* (by adjusting transmitter output), then switch to *reverse* and note the reading. This is the new 2:1 point on the meter and should normally be at one-third scale except for the action of the new series resistor. Suppose this new point reads 70 microamperes on a 100-microamp meter. Multiply 70 times 3 (210) and use this figure as the new full-scale reading for future measurements:

$$\frac{210+70}{210-70}=\frac{280}{140}= 2 \text{ or } 2{:}1 \text{ swr}$$

For a load that results in a reverse reading of 20 microamps:

$$\frac{210+20}{210-20}=\frac{230}{190}= 1.21{:}1 \text{ swr}$$

By juggling the value of the series resistor you can place 2:1 at a given point on the meter up to full scale. Multiply this point by 3 and use this for full scale in the formula.

An alternative method of calibration would be to use 16.6 or 150 ohms for a load and place 3:1 at full scale. This should normally be at half scale so multiply 100 by 2 (200) and proceed as follows:

$$\frac{200+100}{200-100}=\frac{300}{100}= 1.21{:}1 \text{ swr}$$

Of course, as you assign a higher swr to the full-scale position you begin to lose the benefits of this system.

Accurate 50-ohm, 2-watt resistors mounted in PL-259 plugs may be used with the low-power bridge. T fittings may be used to allow one, two or three loads to be put in parallel for 1:1, 2:1 and 3:1 calibration points. In-line meters may be modified in the same way if they are sensitive enough to give full-scale readings on *forward* after installation of the resistor. Original calibration and subsequent tests are made with the meter at full sensitivity and the transmitter output varied for setting to full scale on *forward*. Unless carefully constructed, accuracy of in-line meters will probably not be as good as with the resistance bridge.

This procedure places a calibrated point up on the scale instead of at zero where accuracy is poor. It expands the low end of the meter to give better accuracy for pruning antennas and matching impedances. It pushes *infinity* off the meter face. Within the inherent limitations of these bridges and meter movements this method affords greater accuracy for a very small price.

E. R. Lamprecht, W5NPD

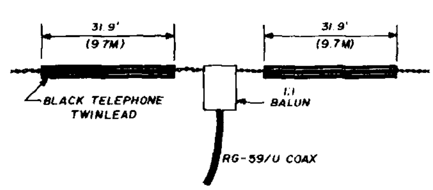


fig. 7. A 40-meter dipole made from black, telephone-company twinlead is shortened slightly because of velocity factor. Swr of this antenna is 1.2:1 at the band edges.

telephone-wire antenna

Recently I built a 40-meter dipole using slightly used, black, two-conductor (twinlead) telephone wire. To broadband the antenna the two wires were connected together at the ends as shown in **fig. 7.** Because of velocity factor I expected that the usual formula for the length of a quarter wavelength wire, $\lambda/4$ *(feet)* = $234/f_{MHz}$* would be too long, and this indeed was the case. After three cuts (three times up and down two trees), a good antenna resulted. The new length formula for this wire, $\lambda/4$ *(feet)* = $229/f_{MHz}$ may save some readers scraped arms and lost tempers. My antenna, which is centered on 7175 kHz, is fed with RG-59/U and has a maximum swr of 1.2:1 at the band edges.

Joel Elston, K9TBD

*A quarter wavelength in metric terms is given by $71.3/f_{MHz}$. The correct formula for the telephone-wire antenna is $69.8/f_{MHz}$ (length of each dipole element in meters).

160-meter shortened vertical antenna

When you're using a grounded, center-loaded vertical antenna and no guy wires are desired, the feed system shown in fig. 8 solves the problem of feeding power to the antenna. The feeder may be inside or outside the lower mast section, but should be kept close to the foot of the mast so its potential is close to ground at that point. The coupling to the center loading coil is made at the mast end or "cold"end of the loading coil. I use three turns of coaxial cable with the center conductor returning to the outer braid at the bottom of the three-turn link.

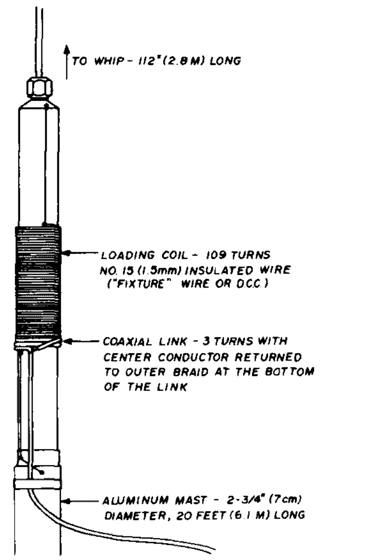
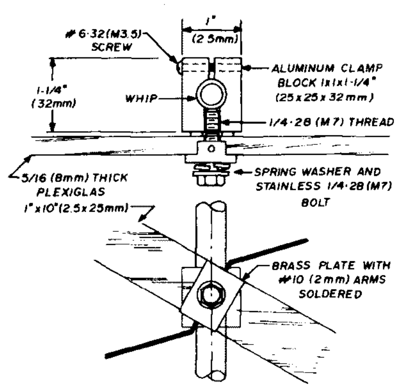


fig. 8. Center loading coil and link coupling used with the shortened vertical antenna for 160 meters. Total antenna height is about 30 feet (9.1m).

My vertical is tuned to 1800 kHz in the 160-meter band and the coil measures approximately 270 μH. This requires a total whip length above the coil of 112 inches (2.84m). The loading coil consists of 109 turns of no.15 (1.5mm) insulated wire, weather proofed with several coats of clear *Krylon* spray. The coil ia 2-3/4 inches (7cm) in diameter and 7-5/8 inches (19.4cm) long. The bottom of the coil is spaced away from the lower mast by about 5 inches (13cm). The lower section of the vertical is made of 2-3/4 inch (7cm) diameter thin-walled aluminum pipe, 20 feet (6.1m) long. This gives a total vertical height of approximately 30 feet (9.1m).

The feed resistance of my vertical, as measured with an Omega noise bridge, is close to 55 ohms. My ground system consists of ground rods and water system as well as a quarter-wavelength of no. 12 (2mm) copper wire just under the surface of the ground.

The bandwidth of the shortened vertical antenna is very narrow so I added the simple capacitive loading system shown in fig. 9. This consists of two 12-inch (30.5cm) lengths of no. 10 (2.6mm) copperweld wire which are attached to a swivel joint. By adjusting the angle of these rods to the mast with halyards (remotely with a selsyn, if desired), it's possible to operate over a 30 kHz bandwidth with a 1:1 vswr.



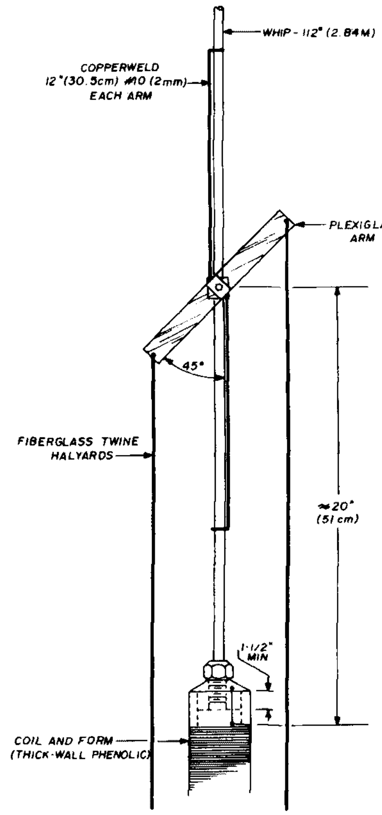


fig. 9. Movable 12" (30.5mm) lengths of no.10 (2mm) copperweld add sufficient capacitive loading that the shortened 160 vertical can be used over a 30 kHz bandwidth with a 1:1 vswr.

Dave Atkins, W6VX

Ham-M modification

The accuracy of the metering circuit in the Ham-M rotator is poor, at best, under conditions of varying line voltage. With a line voltage change from 105 to 125 volts, the full-scale reading in my unit varied from 325° to 365°.

While there are, undoubtedly, many modifications to this circuit that would eliminate this problem, cost and parts availability were a factor. The result, a simple voltage regulator, is shown in the schematic diagram of fig. 10. New parts are given in the parts list. I used two zeners in series, lacking a single one that would render proper performance. With the value shown for R1, the total zener voltage should be somewhere between 17 and 20 volts. The exact value is not critical so long as the voltage across C1 is under the control of the zener when the line voltage drops to 105 V. C1 can be of any value from 500-μF up, depending upon available parts. R2 was added to compensate for the lower voltage across the rotor pot. With the circuit constants shown, there is less than 1° of change from 105 to 125 V input after a 10-minute warm-up.

Walter Pfiester, Jr., W2TQK

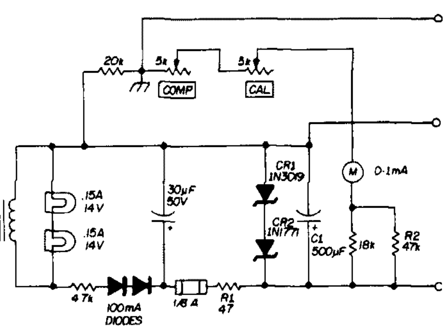
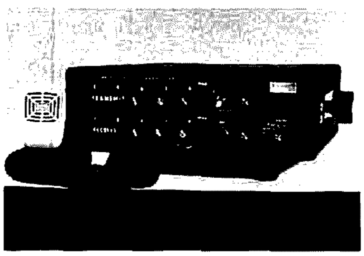


fig. 10. Ham-M meter circuit modification.

NEW products

two-meter fm transceiver



A new arrival on the vhf scene is the Brimstone 144 amateur two-meter fm transceiver by Satan Electronics of Salina, Kansas. Exclusive design features, together with rugged but attractive styling, make this all-solid-state transceiver an outstanding communications package for even the most discriminating user. Frequency generation is by the Satan Electronics "Warlock Frequency Control System," which is a phase-locked-loop synthesizer that provides frequency coverage between 143.00 and 149.99 MHz, with 142-MHz coverage an optional accessory. Frequency selection is in 10-kHz steps by dialing in the desired frequency with rotary selector switches. Or, you can step the frequency in 5-kHz increments simply by pulling out the squelch-control knob. Transmit and receive frequency selection is by separate switches, and you have a choice of either repeater or simplex operation by flipping another switch. The front-panel controls and frequency readout indicators are arranged to provide maximum operating efficiency and convenience. A signal-strength meter gives a clear indication of relative transmitter power output and received signal strength.

The transmitter provides 25 watts nominal power output with a frequency stability of 0.001%, with fm frequency deviation adjustable from zero to 20 kHz. Nonharmonic spurious output is down 80 dB, thanks to the frequency control system. The modulation system uses speech-processed audio applied to a varicap modulator diode. A 500-ohm dynamic microphone is furnished with the transceiver. Any well-designed two-meter antenna will work with the Brimstone 144, which is designed for a nominal 50-ohm load.

The receiver features a low-noise, dual-gate fet rf amplifier with about 18 dB gain and a bandpass filter to minimize image and cross modulation. The receiver is a single-conversion superhet with a 10.7-MHz i-f. Sensitivity is 0.35 microvolt for 20 dB quieting and 0.25 microvolt for 12 dB SINAD. An 8-pole filter in the standard transceiver provides a 2:1 shape factor: ±7.5 and ±15 kHz respectively at 6- and 60-dB bandwidth. Even greater selectivity is available with an optional 12-pole filter.

For added versatility you can choose optional plug-in modules: tone burst for 1800-2400 Hz, Touch-Tone interface, subaudible tone, dial tone, super selectivity, and extended frequency range (142.00-149.99 MHz).

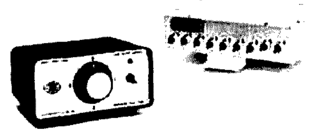
The Brimstone 144 maintenance manual is comprehensive and well written. Large, clear photos of all circuit modules with keyed parts designators are provided, including a list of all parts *and* their manufacturer. The sections on circuit description and maintenance (including troubleshooting) are especially well done, which makes the manual an extremely useful addition to the total Brimstone 144 communications package.

The standard Brimstone 144 transceiver amateur net price is $650.00, which includes the dynamic microphone and a mobile mounting bracket. For an informative brochure, including accessory module prices, write Satan Electronics, Incorporated, 2916 Arnold Avenue, Building 317, Salina, Kansas 67401, or use *check-off* on page 118.

mobile antenna mount

A new approach to mobile antenna mounts for the popular 3/8-inch (9.5mm) blind (one side) installation has been developed by Larsen Electronics, Incorporated. It's the JM antenna mount, which consists of four easy-to -install components: an anchor foot, braid nut, rubber washer, and insulator. The JM mount will accommodate any hf or vhf antenna that adapts to a 5/16-24 (approximately M12) stud. This includes most ¼-*wavelength* (ground plane) and gain-type antennas. Larsen also offers "match-mate" antennas for the JM antenna mount. For more details on the new JM mount, write Larsen Electronics, Incorporated, 11611 Northeast 50th Avenue, P.O. Box 1686, Vancouver, Washington 98663, or use *check-off* on page 118.

beam steering combiner

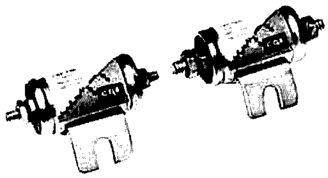


The Omega-t 2000c Beam Steering Combiner provides a low-cost means of beam steering for two- or four-element high-frequency phased arrays. The 2000c is typically used to array vertical monopole or horizontal dipole elements for receiving and transmitting applications, and is useful for arraying any type

More Details? CHECK—OFF Page 118

of elements where increased gain and directivity are required. Matching transformers, power dividers, and delay-line switching are provided for broadband 360-degree beam steering in 30 azimuth steps. Direct dial readout of the selected beam maximum azimuth is provided. Frequency range is 1.8 to 30 MHz, and power rating is 1200 watts PEP or average. For descriptive literature and pricing information, write Electrospace Systems, Inc., 320 Terrace Village, Richardson, Texas 75080, or use *check-off* on page 118.

coaxial feed-through filter



Cornell-Dubilier Electric has added three coaxial feed-through radio-frequency filters to their line of Clear[R] CB noise-filter products. They are the model CBFT 20 (20 amps, 600 working volts dc, 0.1 microfarad); model CBFT 40 (40 amps, 600 working volts dc, 0.5 microfarad); and the model CBFT 60 (60 amps, 50 working volts, 0.5 microfarad).

The CBFT filters are designed to completely enclose the conductor carrying the rf noise component. Since rf travels on the conductor surface, removal of the rf noise component is extremely effective using this type filter. CBFT filters are recommended for use on air conditioners, refrigeration units, voltage regulators, ignition systems and similar equipment. Note the current range available: 20, 40, or 60 amperes — high enough to handle most rf noise problems encountered in industrial equipment. The CBFT filters can also be used in equipment environments found in large tractor-trailer rigs.

For additional information on Cornell-Dubilier's filters, write to William Carlson, Cornell-Dubilier Electric, 150 Avenue L, Newark, New Jersey 07101 or use *check-off* on page 102.

one dollar

ham radio magazine

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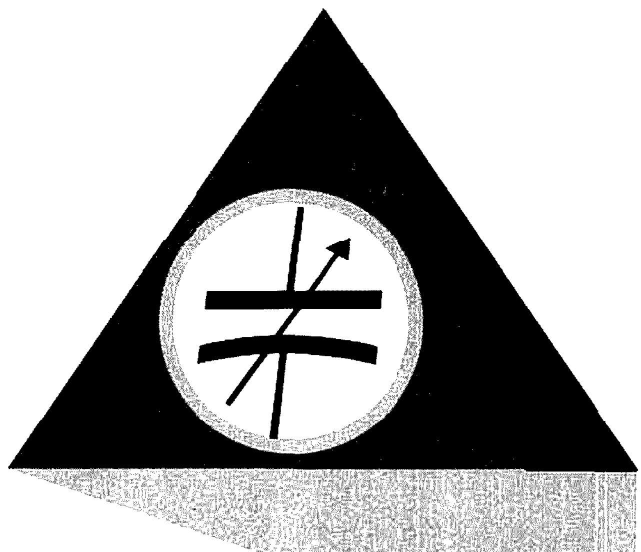

vfo design techniques for improved stability

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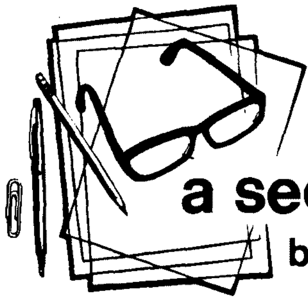


a second look

by Jim Fisk

For the past ten years or so many of the advances in communications circuits and components have been made in the digital area. Although great progress has also occurred in solid-state devices for communications circuits, digital techniques have taken over in many areas previously dominated by more traditional analog circuits – pulse-counting fm detectors, phase-locked loops, frequency synthesizers, high-speed T/R switches, and digital filters are only a few of the more common digital circuits found in modern amateur radio equipment. In the future I expect to see a good deal more innovative digital circuit design as engineers work to improve the operating efficiency and convenience of commercial amateur gear.

When the first low-cost digital circuits became available to amateurs and experimenters in the mid 1960s, the only obvious amateur application was in the electronic keyer. Amateurs soon discovered that frequency calibrators and VOX circuits lended themselves to digital techniques, and digital frequency counters started to replace the venerable BC-221 as the standard for amateur frequency measurement, but other than these somewhat limited applications, digital circuitry seemed to hold little promise for the future of receiver and transmitter design.

However, as the speed of digital circuits increased into the MHz range, and phase-locked loops became commercially available, amateurs were quick to recognize the potential of using these circuits in their own equipment. The equipment manufacturers responded quickly, too, with the result that today's radio gear offers operating accessories and convenience not previously available to the amateur operator. Complex commercial communications circuits that were once considered far too expensive for the amateur market are now commonplace, and recently introduced equipment offers more performance per dollar than most amateurs ever thought possible.

The new Kenwood TS-820 is a prime example of innovative design in amateur radio gear. In addition to such niceties as i-f passband tuning, rf speech processing, rf monitor, rf attenuator, and digital readout (optional plug-in accessory), the TS-820 features the use of a phase-lock filter in the local-oscillator circuit, a first in commercial amateur radio equipment. Although most amateurs have been exposed to the use of the phase-locked loop in frequency synthesizers, this is not its only function. Its original use, in fact, was to filter very noisy signals, and that's how it's used in the new Kenwood transceiver.

In most receivers and transceivers the vfo is mixed with a crystal oscillator signal to provide the required transmit-receive frequency – the vfo always tunes the same low-frequency range to maintain overall stability. The problem with this approach is that, instead of a single injection signal, there are at least four signals present (the vfo, the crystal oscillator, and the sum and difference of both). Since these additional signals cause spurious responses in both the receiver and transmitter, overall operation can be greatly improved if a clean injection signal is available. The phase-locked source is able to suppress all harmonics of the reference frequency by as much as 120 dB – this is inherent in the basic phase-lock circuit and doesn't require any additional filtering.

In the TS-820 the signals from the vfo, crystal oscillator, and carrier oscillator are combined in a phase detector to produce a control signal which sets the frequency and phase of a voltage-controlled oscillator that provides the actual injection signal. The result is a clean sinusoidal injection signal with all control signals attenuated by more than 100 dB. This is the key to the TS-820's very high spurious rejection specification of more than 80 dB.

Although Kenwood is the first to apply this important technique to an amateur transceiver, it is indicative of the innovative thinking of many amateur equipment manufacturers. In the future I expect that amateur equipment designers, inspired by low-cost, high-performance digital circuits, will be able to offer equipment performance and convenience not now available at any price.

Jim Fisk, W1DTY
editor-in-chief

presstop

EXTRA CLASS LICENSEES who've been on the air at least 25 years will be the first to have their choice of open callsigns as a result of FCC's Report and Order on Docket 20092 released in late April. The Report and Order, which becomes effective July 1, gives first priority to Extra Class licensees already eligible for a one-by-two (e.g., K1AA) callsign for a three-month period starting July 1. Then, after all those eligible during the first period have exercised their options (if they wished to) a second group, which includes all additional Extra Class licensees who were licensed as Extra Class by November 22, 1967, will be eligible to apply for the remaining callsigns of their choice.
Two Additional three-month periods, the first for all who received their Extras on or before July 2, 1974 and the final for those who were Extras by July 1, 1976, follow. After that, presumedly, any Extra Class licensee, no matter how short a time he's been licensed, will be able to apply for the available callsign of his choice.

LONG AWAITED "BANDWIDTH" Notice of Proposed Rule Making has also been approved by the Commissioners and proposes changing the present sub-band divisions by emission type to divisions predicated on maximum permissible bandwidth. Under the terms of the NPRM any type of emission would be permitted on a given frequency as long as it did not exceed the bandwidth limits for that frequency.
The Three Bandwidths proposed in this new and revolutionary approach to dividing up the Amateur sub-bands are 350 Hz, 3.5 kHz and 35 kHz, presumedly equivalent to the present CW, SSB and FM divisions. It's definitely a "deregulatory" proposal, permitting as it does experimentation with new techniques while encouraging spectrum conservation. It also has some serious potential problem areas such as already exist between slow-scan TV users and SSB DXers on the low end of the 20-meter phone band.

RADIO DEALERS MODIFYING Amateur Radio transmitting equipment to operate on CB frequencies are treading on very thin ice, according to a recent conversation with the FCC. Making such a conversion appears to be a violation of Section 302 of the Communications Act since the conversion makes the equipment a piece of non-Type Accepted CB equipment. It probably also puts the dealer in a "Catch 22" type situation — if he has a commercial license he has a good chance of losing it, and if he doesn't he is subject to penalty for working on "commercial" equipment without one.

FCC STUDY GUIDES for both Novice Class and General/Conditional/Technician Class Amateur exams were released by the FCC in April. Though they are very general in scope the guides would be a useful tool for anyone planning to take an FCC Amateur exam.
Copies Of The Study Guides are available from Ham Radio. Send an SASE with 24¢ postage for each guide and be sure to specify which you want.

FCC FORM 610, the application for an Amateur license, is not currently available from the FCC in Washington. Their supply is exhausted and won't be replenished until a revised Form 610 that is currently being developed receives final OK and can be printed — and that's probably a month from now!
Both Ham Radio And ARRL do have some stocks of Form 610 and Ham Radio plans another printing "just in case." A photocopy or other reproduction of Form 610 may also be used — the only stipulation is that reproductions must be the same size as the FCC original.

WRITTEN CW TESTS are likely to be with us for some time to come despite the recent relaxation of the rules to permit code "comprehension" exams. The new procedures still haven't been firmed up, and even after they are it will take a while before new tapes and matching examinations can be produced and distributed. The change should take place some time this year, though — late fall seems likely.

PC-76, THIS YEAR'S "PERSONAL COMMUNICATIONS" TRADE SHOW in Las Vegas, while primarily CB, did have some interesting new items for the many Amateurs who attended. Most exciting was probably Hy-Gain's new all-band (10-160) 200-watt transceiver, which boasts a number of neat features like an LED digital readout with memory for storing and recalling an interesting frequency for later reference. A new entry to two meters is Fieldmaster, which is importing the Multi-2000 and Multi-11 from Japan, and CEPCO had touchtone pads and decoders for Amateur autopatch use. Breaker was showing a VHF mobile antenna line they'll be bringing out this summer under the Hallicrafters name, and Silitronix had the Swan SS-747 in its booth. Henry introduced three new VHF/UHF pocket receivers to complement its popular MR-2, and both Midland and TPL had their Amateur equipment on display. ARRL's booth had a steady stream of visitors, both individual Amateurs and distributers interested in getting in on the forthcoming Amateur Radio boom.

vfo design techniques

for improved stability

Complete description of a new and improved solid-state variable-frequency oscillator with exceptional frequency stability

Baffled by the pros and cons of solid-state vfo design? A great many points of view are given in published material concerning vfos, so it's a small wonder that a newcomer to solid-state design finds himself in a state of perplexity when he attempts his first variable-frequency oscillator design. Some amateur writers laud the Franklin vfo circuit, while others express a definite preference for the Seiler or Vackar designs. Each type of oscillator has keynote qualities – even the common Colpitts variety. It can be said without reservation that each in the foregoing list can be made adequately stable for most amateur work by following a few simple design rules. Some guidelines are offered here. They are based on considerable laboratory work, and represent the cream of data which I gathered during the course of three years of investigation.

circuit choice

Having been exposed for many years to tube-type vfos with their confounding tendencies to drift and produce hum-modulated output voltage, the appearance of transistors was a welcome event, indeed. However, It was learned that a significant reduction in heat through the application of semiconductor active devices did not always result in acceptable vfo stability. Furthermore, experience proved that a considerable amount of frequency drift resulted from rf-current heating within the capacitors, transistors, and inductors used in a vfo. Such heating causes changes in component values, and drift prevails. Elimination of ac voltage practically resolved the hum-modulation syndrome in most transistor vfos, as there were no filaments to heat. This advantage also reduced the long-term drift problems related to tube-type variable-frequency oscillators.

Many experimenters, and I was among them, chose the parallel-tuned Colpitts as a practical oscillator. In fact, most commercial designers of amateur equipment still use that breed of oscillator with fets and bipolar transistors. Various stability profiles have resulted from the use of Colpitts vfos, but I have yet to see one – commercial or home designed – that exhibits a long-term drift characteristic better than 60 or 70 Hz from a cold start to some period a few hours later. Worst-case frequency runs have unmasked solid-state oscillators which drifted as much as 2 kHz before stabilizing. After "stability" was attained, it was not unusual to find the oscillator rambling up and down constantly over a 100-Hz range. This characteristic can be distressing,

By **Doug DeMaw, W1CER**, Technical Editor, *QST Magazine*, 225 Main Street, Newington, Connecticut 06111

especially when the operator finds it necessary to keep readjusting the operating frequency during a QSO. Matters are complicated further when narrow-band i-f filters are employed, say, a 300-Hz passband for *CW* work.

I needed a better vfo than could be provided easily by a simple parallel-tuned Colpitts. At some time during the course of numerous experiments there came an important recollection from vacuum-tube days: best stability always seemed to result when the then-popular series-tuned Clapp vfo was used. It was reasonable to conclude that the principle was applicable in solid-state design work as well. An incredible improvement in long- and short-term drift was noted when tests were performed

necting leads related to the coil degrade the tuned-circuit Q by virtue of being parasitic inductances in series with the desired inductance. The foregoing illustrates the undesirable aspects of a high-C, low-L oscillator tank.

In a typical series-tuned vfo tank, an inductance of 4 μH might be required with the same amount of tuned-circuit capacitance, thereby greatly reducing the bad effects just outlined. In an ideal vfo one would not use a coil with a magnetic core. Rather, an air-dielectric inductor of large wire diameter and superb rigidity would be employed. This approach would impose certain physical restrictions which would not be acceptable to many amateur builders — notably those who specialize in assembling compact equipment.

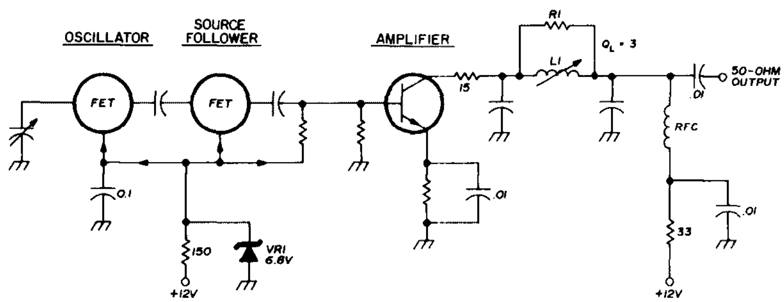


fig. 1. Diagram showing regulated voltage applied to various parts of a vfo circuit. A 50-ohm output pi-network tuned circuit is used at the amplifier output.

on the first model built. A succession of circuit improvements followed, each providing greater insight into the causes of drift. Those matters are treated here.

observations

A major cause of drift in a bipolar oscillator which contains a parallel-tuned LC circuit is inductance change. Typically, an inductor is used which relies on powdered iron or ferrite as a core material. Slug-tuned coils are common choices because they enable the designer to take advantage of the small size -- a product of our present trend toward miniaturization. However, most bipolar oscillators require a high C to L ratio, resulting from the typically low impedance levels of bipolar-device input ports. Because of the foregoing, a 40-meter vfo, for example, might have a total tuning capacitance of 500 pF at its lowest operating frequency, and an inductance of only 1 μH. In this situation we have an inductor which is subject to a high percentage of change when the total inductance varies by only a fraction of a microhenry. Such a change, however slight, will cause a major shift in operating frequency. The change will probably result from slight variations in the core properties of the inductor.

Additionally, mechanical stability of such a circuit is borderline because the leads going to and from the inductor, plus the PC-board elements which are common to that part of the circuit, comprise a significant portion of the inductance. Flexing of the vfo circuit board or chassis can shift the operating frequency markedly. Another disadvantage of a low-L circuit is that the con-

A good practice is to use a slug-tuned coil which has a low-permeability core, and that core should just enter the coil winding at resonance rather than occupy most of the area within the coil. This will minimize thermal drift caused by changes in core properties. Powdered-iron cores are more stable than ferrite ones, and are highly recommended when a core must be used. The slug mechanism should be physically stable when in the coil-form collar. This will help prevent frequency changes caused by slight movement of the mechanical components of the coil assembly.

Perhaps one of the worst cores you can use when striving for optimum stability is a toroid. Changes in core characteristics are quite pronounced as the ambient temperature varies. For casual design work it is possible to get satisfactory results with powdered-iron toroids, but they are best suited to environments where the room temperature is relatively constant. A heavy coating of Q-dope on a toroid coil will enhance stability because the cement will keep the coil winding from shifting position. In fact, doping the windings of slug-tuned coils is similarly beneficial. Ceramic or steatite slug-tuned coil forms are best for vfo work, and phenolic material should be used only when nothing else is available.

It has been more or less standard procedure for amateurs to use silver-mica capacitors in the frequency-determining part of a vfo. The stability characteristics of these units are pretty good, but seldom good enough. That is, the drift characteristics from a given production run of these capacitors may vary considerably when they are placed in an oscillator circuit and checked one by

one. I am referring here to temperature tolerance versus rf heating and changes in ambient temperature. Ceramic compensating capacitors can often be added to a vfo which contains silver micas, and eventually a combination can be found which provides acceptable stability. However, this exercise can become one of frustration which most amateurs would like to avoid.

Some experiments with polystyrene capacitors showed a remarkable reduction in nonuniformity with respect to changes from heat. Whenever these low-cost capacitors were used, drift was so minor that compensating capacitors could be eliminated if ideal stability was not a prime criterion. Suddenly, the path to stability was becoming shorter, and I adopted polystyrene capacitors for all succeeding vfos.* The results were always superior to those obtained with plain or dipped silver micas.

Another trial-and-error performance standard developed while comparing the drift traits of various trimmer capacitors. It was found that the worst of the lot for vfo applications was the ceramic trimmer. Not only did changes occur in the preset capacitance when ambient temperature excursions took place, but also changes in heat caused additional drift as a result of mechanical shifts in the movable parts of the trimmers. Among the best trimmers I have used are the E. F. Johnson silver-plated, PC-board, air trimmers. The plates are milled rather than manufactured separately. This results in a unit with superb mechanical stability. The Q of these trimmers is also very high.

A further contribution to oscillator stability can be realized through application of moderate operating voltages. That is, if a 12-volt dc supply is available it's better to lower the drain or collector voltage to 6.8 or some other low potential in that general area. Of course, the addition of a zener diode to regulate the reduced voltage is recommended. The reduced operating voltage lowers the oscillator power, which as a result reduces heating and changes in junction capacitance. It is best to develop the required oscillator-chain power *after* the vfo stage by means of class-A buffer/amplifier stages.

Finally, in the long search for improved stability traits, a technique introduced in *QST* by George Hanchett, W2YM, was tested and adopted in all vfos which employed jfets and mosfets[1]. This calls for the addition of a high-speed silicon switching diode from the oscillator gate to ground. The anode is connected to the gate terminal of the device. The principle is one of regulating the transistor bias, which of course enhances stability. Furthermore, the diode clamps on the positive-going sine wave to limit the transconductance, which in turn minimizes changes in junction capacitance. This feature greatly reduces the harmonic currents in the oscillator output – another good design goal. A 1N914A or equivalent diode is satisfactory for the purpose. The beneficial effects are most pronounced when a source-bias resistor is used in the oscillator. If the source return has no appreciable dc resistance, the gate-source junction of the fet tends to perform the same function as the diode, but not as completely. Therefore, some improvement is always obtained by adding the diode.

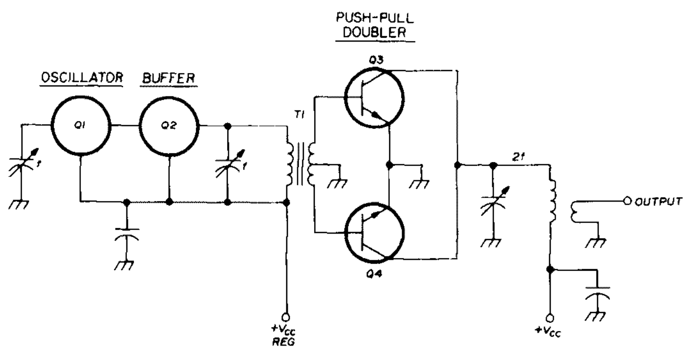


fig. 2. Illustration of a basic push-push class-C doubler.

isolating the load

Stability is not totally dependent upon the oscillator parameters. Changes in the load to which the vfo is connected will cause phase shifts which change the oscillator frequency. The condition is seen when studying a chirpy CW signal. Of course, unwanted rf which is allowed to get into the vfo circuits can cause a similar condition, but we shall ignore that matter here.

Because of the effect load changes bring about, the lightest amount of coupling practicable should be used between the oscillator output and the stage to which it is connected. It is not recommended that series resistance be placed in the signal line (in series with the output voltage), even though this is done by some designers to reduce the coupling. The addition of series resistance contributes to vfo noise output, which is a very undesirable characteristic: a well-designed vfo should have a noise plateau 90 dB or more below the desired energy output level. This is especially true when the vfo is used in a receiver. For this reason it is best to use capacitive coupling from the oscillator, making the capacitor as small in value as can be tolerated. Additionally, the succeeding buffer stage should represent a fairly high input impedance to lessen oscillator loading. An fet

*Based on ARRL laboratory findings, polystyrene capacitors are suitable for applications from dc to at least 30 MHz

source follower makes a fine buffer stage in this case. Polystyrene coupling capacitors are recommended between the vfo and buffer stages, as changes in capacitance due to heating will cause minor changes in the operating frequency of an oscillator.

I have always obtained the best results by having two isolating stages after the oscillator. An fet source follower with a broadly resonant source circuit will increase the signal level to the second buffer, and this is desirable when very light oscillator coupling is used. The second buffer can be a bipolar transistor operating in class A. This will amplify the signal to a practical level for most amateur work, and will add to the isolating properties of the overall circuit. When sufficient power is available from the source follower, the last stage (buffer/amplifier) may be operated in class C, but this will increase the harmonic currents in the collector circuit. Forward bias for the bipolar amplifier should be obtained by means of the usual resistive divider, but the voltage source for the divider should be a regulated one – typically the same one used to power the oscillator. This practice will also help reduce load changes. An example of the principle is shown in **fig. 1**.

Shown also in **fig. 1** are the details of a recommended output network for minimizing oscillator pulling from load changes. The rationale calls for a low-impedance output port . . . in the region of 50 ohms. The goal is not one of impedance matching. Rather, it is to develop a low characteristic output impedance which will be relatively immune to external load changes. In most practical circuits the transmitter stage which follows the oscillator chain is operated in class A, and uses a bipolar transistor. This puts the input resistance of the outboard stage in the region of 500 to 1000 ohms for low-level amplification. Since the characteristic impedance is somewhat higher than that of the last vfo buffer, changes in level have a minor effect on the oscillator chain.

L1, C1 and C2 comprise a pi network in **fig. 1**. This increases the output voltage and discriminates against harmonics. To ensure ample bandwidth for the vfo tuning range a low-Q network is employed. Furthermore, it is helpful to shunt L1 with a swamping resistor – usually 3300 to 5600 ohms. If the output transistor has a high beta, some instability can occur when the oscillator chain is connected to a high-impedance load. It's the usual open-loop gain syndrome versus unconditional stability. Therefore, R1 assures stable operation even when no load is connected to the overall vfo circuit. R1 also increases the bandwidth of the buffer/amplifier.

mechanical considerations

A generally accepted practice these days is to use double-sided PC board in solid-state work. The back side of such boards is used as a ground plane to minimize rf current loops and to discourage vhf and uhf parasitic oscillations. The technique is a good one indeed, but it can greatly complicate a vfo design if double-sided board is used in the immediate vicinity of the oscillator. The etched circuit-board elements function in combination with the ground plane to form numerous low-value capacitors, and the epoxy or phenolic insulating material of the PC board becomes the dielectric element of the capacitors. Capacitors of this kind are low in Q and can degrade the oscillator tuned circuit. Worse yet, capacitors formed in that manner are extremely subject to changes in value as variations in ambient temperature occur. As a general rule, therefore, no part of the oscillator chain should be built on double-clad board, just to stay on the safe side of things.

The entire vfo assembly (buffers included) should be housed in a shielded enclosure. This will help to reduce long-term drift and will offer isolation against stray oscillator radiation. The technique will also prevent rf from circuits external to the vfo from entering the oscillator chain. The shield-box walls should be quite rigid to prevent them from moving toward or away from the tuned-circuit elements of the oscillator as this would cause changes in frequency.

The vfo main-tuning capacitor should be of the

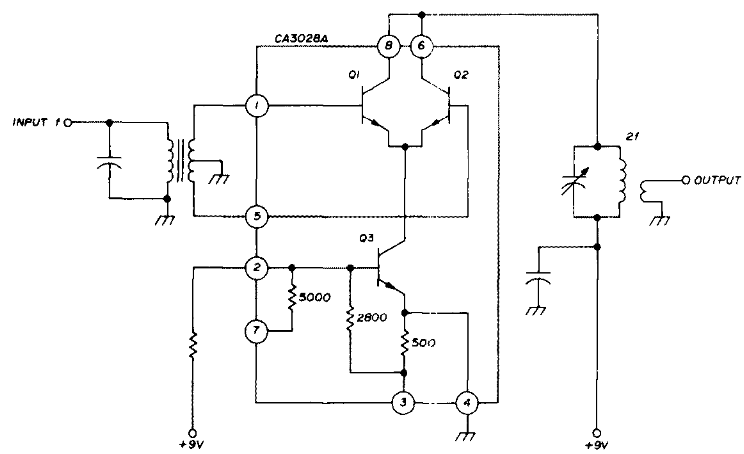


fig. 3. A method for using the matched characteristics of a CA3028A IC in a push-push doubler, as suggested by W7ZOI.

double-bearing variety. It should have a shaft which turns with minimum torque, and should be driven by a smooth-running reduction mechanism. There should be no significant stress required when changing the operating frequency from the front panel of the equipment. Also, the rotor of the tuning capacitor should be grounded at both ends of the frame. Variable capacitors with aluminum plates have a greater drift characteristic than those with plated brass vanes. I personally prefer the latter.

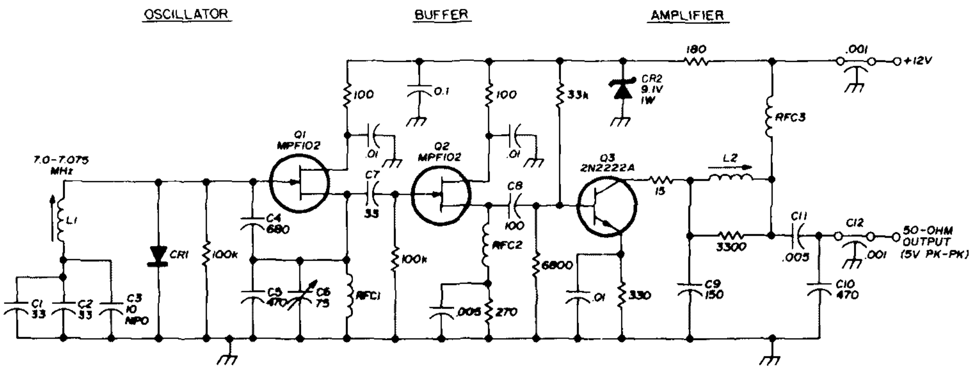
another isolation technique

It was learned many years ago by tube-oriented amateurs that oscillator pulling from load changes could be reduced significantly by operating the oscillator one octave lower than the desired output frequency. The principle has merit, and is completely germane to solid-state design work. It was not uncommon to find amateurs using a 40-meter vfo which employed a 3.5-MHz oscillator. Of course, a frequency doubler had to be included in the oscillator chain, but most designs used a single-ended multiplier.

Better multiplier efficiency can be obtained by using a push-push doubler, as it has an efficiency profile nearly equivalent to that of a straight class-C amplifier. **Fig. 2** shows a circuit example of a push-push doubler. It can be seen that the drive to Q3 and Q4 is supplied in push pull by means of T1. However, the collectors of the class-C transistor pair are in parallel. The collector tuned circuit is resonant at twice the drive frequency. If good dynamic balance is effected at Q3 and Q4, the output waveform will be nearly a pure sine wave at the desired frequency. The example does not show the necessary balancing controls for discrete devices. A practical vfo circuit which includes the required components is presented later in this paper.

Dynamic balance is important in a push-push doubler to ensure that a minimum amount of the driving frequency appears at the doubler output port. In laboratory experiments it was learned that a well balanced push-push doubler yielded a clean output waveform when a 1000-ohm resistor was used in place of the tuned circuit. However, because of the *IR* drop imposed by the resistance there was a marked reduction in doubler output power.

An interesting approach to effecting dynamic balance was suggested to me during an exchange of correspondence with W7ZOI. He offered that the use of a CA3028A IC would practically eliminate the need for balancing controls in a push-push doubler. This is because the bipolar transistors on the IC substrate have nearly identical characteristics, because they are manufactured from the same silicon slice in a given production run. **Fig. 3** shows how a CA3028A or similar IC might be configured to function as a doubler. The resistances shown inside the rectangle are those which are part of the IC. Terminal 4 has been grounded in this example, and a 2200-ohm resistor has been placed in the base lead of the current source, Q3. In this arrangement Q3 is saturated, bringing the emitters of differential pair



X_{L1} 265 ohms	X_{C3} 2275 ohms	X_{C7} 690 ohms	X_{C11} 4.5 ohms
X_{L2} 140 ohms	X_{C4} 33 ohms	X_{C8} 227 ohms	X_{C12} 23 ohms
X_{C1} 690 ohms	X_{C5} 48 ohms	X_{C9} 152 ohms	X_{RFC1} 4400 ohms
X_{C2} 690 ohms	X_{C6} 303 ohms (minimum)	X_{C10} 48 ohms	X_{RFC2} 4400 ohms

C1,C2,C4,	polystyrene capacitor (Mallory type SX or equivalent)
C5, C6, C7, C8	75 pF miniature air variable, double-bearing type (one section of a Miller 2109 suitable)
C9,C10	dipped silver-mica capacitor
CR1	high-speed silicon switching diode (1N914A or equivalent)
L1	variable inductor, 6 μH nominal (Miller 42A686CBI)
L2	variable inductor, 3.2 μH nominal (Miller 23A47RPC)
RFC1,RFC2	100 μH miniature rf choke (Millen J301 type)
RFC3	500 μH miniature rf choke (Millen J301 type)
Z1	9.1-volt, 1-watt zener diode

fig. 4. Schematic diagram of a practical 7-MHz vfo. Fixed-value capacitors are disk ceramic unless otherwise indicated. Resistors are 1/2 watt composition.

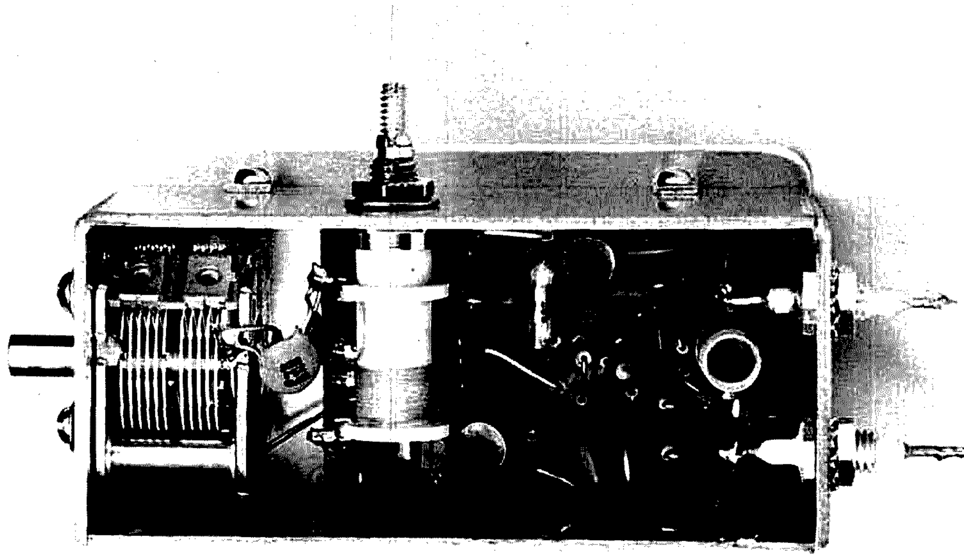


fig. 5. View of the assembled 40-meter vfo. The silver-mica capacitor at the right of the large coil was replaced by a polystyrene unit, as was the one between the coil and tuning capacitor. The vfo shield box is made from sections of double-sided PC board. When in use, an aluminum top cover is pressed in place (friction fit) to completely enclose the circuit.

Q1 and Q2 to dc and rf ground. The differential pair of transistors now operates in class C. A test of this circuit indicated excellent dynamic balance without any need for external tweaking controls.

some practical vfos

Fig. 4 illustrates the circuit for a practical 40-meter vfo chain. The circuit can be scaled to other operating frequencies in the medium- and high-frequency spectrum by using the X_L and X_C values listed to compute the coil and capacitor values. The results will bring the circuit into a workable area. Final adjustments for frequency and bandspread can be done experimentally.

Maximum drift with this circuit (cold start to a period three hours later, at 25°C) was 25 Hz. Only 30 seconds were required to reach the point of maximum frequency change (25 Hz), and the major part of that change resulted from junction heating at Q1, plus the slight heating within the coil and capacitors of the gate tank, caused by the flow of rf current. After stabilization occurred, a "hunting" of no more than 5 Hz was noted. RFC1 and RFC2 are broadly self-resonant at 7 MHz with the existing stray circuit capacitance.

Three series capacitors are used at the lower end of L1. It was learned during lab experiments that drift could be reduced considerably by paralleling two or more capacitors in that part of the circuit. The improvement comes from a division of rf-current flow among several capacitors rather than one. Therefore, the heating effects on any one capacitor are greatly reduced. In general terms, the same technique applies to the feedback capacitors, C4 and C5.

The output energy from this circuit is quite clean. The noise is at a low enough level (−90 dB or greater) to be beyond meaningful measurement with a basic spectrum analyzer setup. The second harmonic was down in excess of 36 dB, and the third harmonic was 45 dB below the fundamental output. Greater spectral purity can be had by inserting a 50-ohm half-wave filter in the output line. However, for most amateur applications it is unnecessary to sanitize the output waveform to that extent.

It should be noted that C12 of fig. 4 is a 0.001-μF feedthrough type capacitor. The combined value of C10, C11, and C12 serve as the output capacitance of the pi-section tank. In this arrangement the feedthrough capacitor is used also as a feedthrough terminal on the shield enclosure. The details can be seen in the photograph (fig. 5).

Immunity to load changes was checked by placing a dead short across the circuit output at C12. The frequency change was noted by means of a counter. The shift from an open circuit to a dead short was only 40 Hz. The pi network was modified to have a characteristic

output impedance of 1000 ohms. From an open to a shorted condition the frequency changes almost 400 Hz. This indicates the value of having a low output impedance in an oscillator chain.

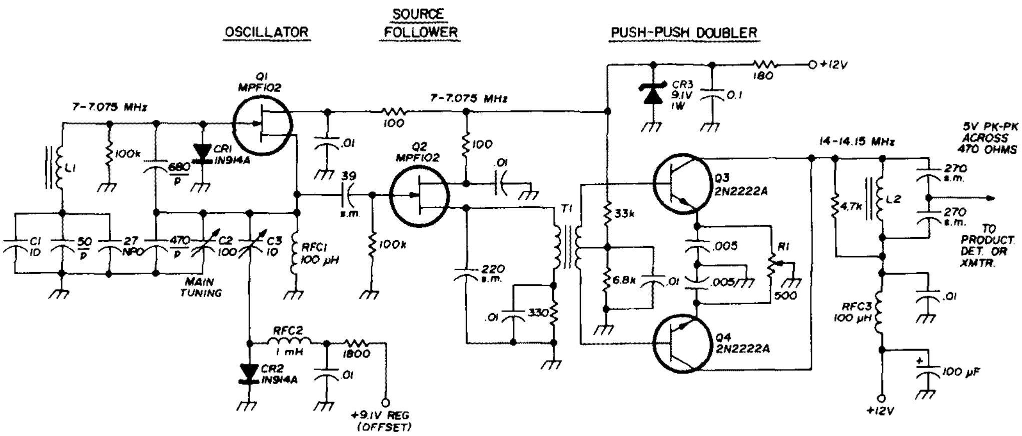
practical push-push doubler

An oscillator chain was needed for use with a QRP transceiver which contained a direct-conversion receiver and a straight-through transmitter strip. The operating frequency of the transceiver was 14 MHz. Rather than tackle the problem of instability at 14 MHz, it was decided that a 7-MHz vfo would be used with a push-push doubler. Also, such an arrangement would minimize the occasion for chirp during CW transmit periods. The circuit under discussion is illustrated in fig. 6.

The oscillator is similar to that of fig. 4. However, a tuned toroidal tank is used in the source of the intermediate stage, Q2, rather than an rf choke, as was done in fig. 4. T1 is necessary in order to furnish drive to the doubler in push-pull. A center-tapped secondary winding accomplishes that. Forward bias is applied to Q3 and Q4 to make the doubler easier to drive. In operation, the transistors are driven into the class-C region and the efficiency is good.

Since discrete devices are used at Q3 and Q4, some means for establishing dynamic balance is required. For that purpose R1 has been included in the circuit. It is adjusted while the circuit is operating, and is set at a point which provides best output-waveform purity. Examination of the output energy was done while using a Tektronix 453 scope (50-MHz bandwidth). There was no visual evidence of 7-MHz or harmonic energy on the waveform after R1 was properly set.

The characteristic impedance of the output port of this oscillator chain is approximately 500 ohms. The lower 50-ohm output condition recommended earlier is not necessary in this design because of the excellent isolation afforded by the doubler. The 4700-ohm swamping resistor across L2 is used to broaden the response of the tuned circuit to permit uniform voltage output from 14 to 14.15 MHz.



C1,C3	subminiature PC-board-mount air trimmers (Johnson 189-507-5 or T-9-5)	R1	500-ohm PC-board-mount control
C2	100 pF miniature air variable (large gang of Miller 2109 suitable)	T1	toroidal transformer. Primary, 2 µH. Use 23 turns no. 24 (0.5mm) enamelled wire on Amidon T-50-6 toroidal core. Secondary is 20 turns no. 24 (0.5mm) enamelled wire (center tapped) over primary winding. Observe same rotation sense when winding
L1	7.6 µH toroidal inductor. 37 turns no. 24 (0.5mm) enamelled wire on Amidon T-68-2 toroidal core (see text)		
L2	1 µH toroidal inductor. 14 turns no. 22 (0.6mm) enamelled wire on Amidon T-50-6 toroidal core	Z1	9.1 volt, 1 watt, zener-diode regulator

fig. 6. Schematic diagram of the vfo and doubler for 20-meter use. Fixed value capacitors are disk ceramic unless noted otherwise. P = polystyrene. SM = silver mica. Fixed-value resistors are 1/2-watt composition.

Fig. 7 shows a breadboard version of the circuit on a test fixture with a direct-conversion receiver. A toroidal inductor was tried at L1, and once doped with cement provided acceptable stability in a near-constant temperature environment. The circuit would be entirely suitable for most home-station applications, but is not recommended for portable use where temperature excursions are the rule rather than the exception. The inductor specified in fig. 4 is better suited for general use of the variable-frequency oscillator.

Stability tests were run with a slug-tuned coil at L1, but without a shield box around the vfo assembly. The drift characteristics were similar to those of the vfo in fig. 4, but the change was multiplied by a factor of two because of the frequency doubler, Q3 and Q4. Total drift over a three-hour period was roughly 70 Hz. Slightly greater "hunting" took place after stabilization because of slight air currents in the room. The hunting never exceeded 15 Hz, however.

When a dead short was placed across the doubler output port, a maximum frequency change of 15 Hz was

observed. This illustrates the good isolation properties of a frequency multiplier when used in a vfo chain. No sign of chirp could be detected on the transmitter signal while monitoring the CW with a receiver set for a 300-Hz i-f bandwidth.

An offset circuit is shown in the oscillator of fig. 6. CR2 is used as a diode switch to offset the vfo frequency by approximately 700 Hz during receive (necessary with direct-conversion transceivers). C3 is set to provide the required offset amount while CR2 is saturated by application of 9.1 volts regulated (available at Z1). C1 is used to calibrate the vfo if a toroidal inductor is used at L1. It need not be included if a slug-tuned coil is employed.

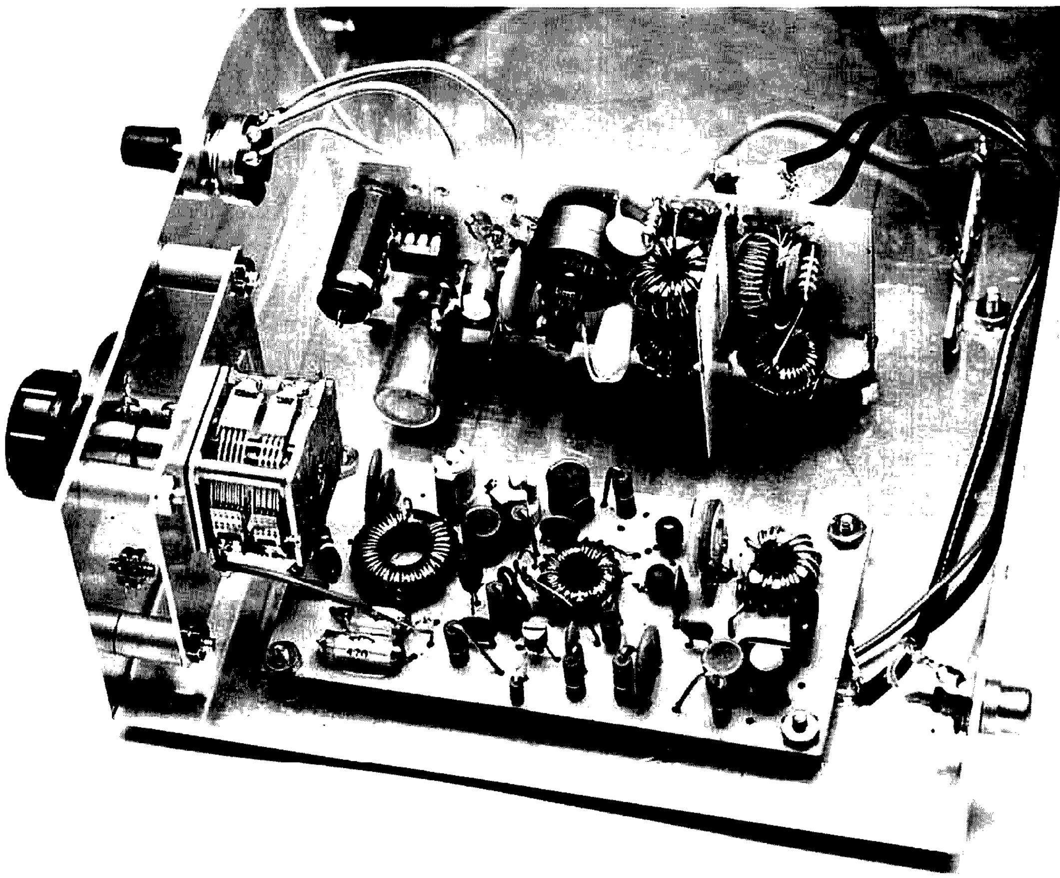


fig. 7. Photograph of the vfo and doubler, as installed in a test fixture. The module in the foreground is the vfo. A direct-conversion receiver is mounted vertically behind the vfo.

summary comments

A 160-meter version of the vfo shown in fig. 4 was built for use in a 10-watt CW transmitter.[2] No drift could be measured with a frequency counter. The last digit of the counter rambled up and down 1 Hz, but that was the maximum amount of change noted during 15 individual attempts to measure the drift over a period of several days. It is entirely possible that through chance luck an exceptional group of capacitors was installed in the oscillator the first time around. But, when the circuit was duplicated later the same results were noted.

All of the vfos discussed here were subjected to direct keying by breaking the supply voltage line to the oscillator. In each instance a chirpless CW note was obtained, but with some clicks which resulted from not employing a shaping network.

Perhaps the basic guidelines offered here will be of value to amateurs who enjoy building their own solid-state equipment. The information should supply the basis for innovation, and who knows? Someone may develop a 7-MHz vfo that is as stable as the one for 160-meters discussed in the foregoing text.

references

1. G. D. Hanchett, W2YM, "The Field-Effect Transistor as a Stable VFO Element," *QST*, December, 1966, page 11.
2. D. DeMaw, W1CER, "More Basics on Solid-State Transmitter Design," *QST*, November, 1974, page 22.

ham radio

time/date printout for RTTY

Multiplexer ICs and a PROM are combined in a circuit that transmits ID information in a 32-character format

The time printout is becoming popular with RTTY operators as a 10-minute ID check and as a useful addition for auto-control stations. This article shows one way to make your own programmable stunt box with enough variations to allow customization to fit your own preferences. The unit prints out time, date, and 16 characters. My unit prints out UNIVERSAL TIME 2146-02/12/75, as an example.

The heart of the unit is a 256-bit programmable read-only memory (Monolithic Memories 6330 or Signetics 8223). This PROM is the equivalent of a 256-diode matrix (32 x 8) in a 16-pin dual in-line package. Not included in the article is the digital clock proper as many such circuits are available; however, only those clocks with TTL levels and BCD format can be used. For simplification, the logic is designed to transmit two full stop pulses (8.0).

circuit description

The schematic is in two parts, timing and control (fig. 1) and information transfer (fig. 3). The output message is thirty-two 8-bit words. A word is one RTTY character, consisting of 1 space, 5 information, and 2 stop bits. Sixteen characters are directly read out of the PROM. They may be any message/operation and may occur before or after the time readout. The other 16 characters are read from a 4-pole 16-position multiplexer and may be programmed for the desired time format.

timing and control

Integrated circuits U1-U4, the oscillator and countdown chain, provide the timing pulses for the baudot divider, ICs U5-U8. When S1 is in the 60 position (fig. 1) U5, a 4-pole, 2-position multiplexer, switches to divide 1 kHz by 22 to obtain the 45.45 Hz baud rate (22 ms pulse width) for 60 wpm. When S1 is in the 100 position, the multiplexer switches and sets up to divide 10 kHz by 132 to obtain the 75.75 Hz baud rate (13.2 ms pulse width) for 100 wpm. U9, U10 provide the binary timing sequencing of the serial output and are held in reset when the circuit is nonoperating. U11A U11B are a flip-flop that starts the circuit functioning when S2 is depressed.

Relay driver Q1 turns on during operation and energizes relay K. Relay K contacts, placed across your loop transistor emitter-base terminals, turn off the loop transistor during the time-sending operation. U11C gates the serial time pulses, which are inverted by U11D, and

By Don Purland, WØLZT, 8201 Northwood Parkway, New Hope, Minnesota 55427

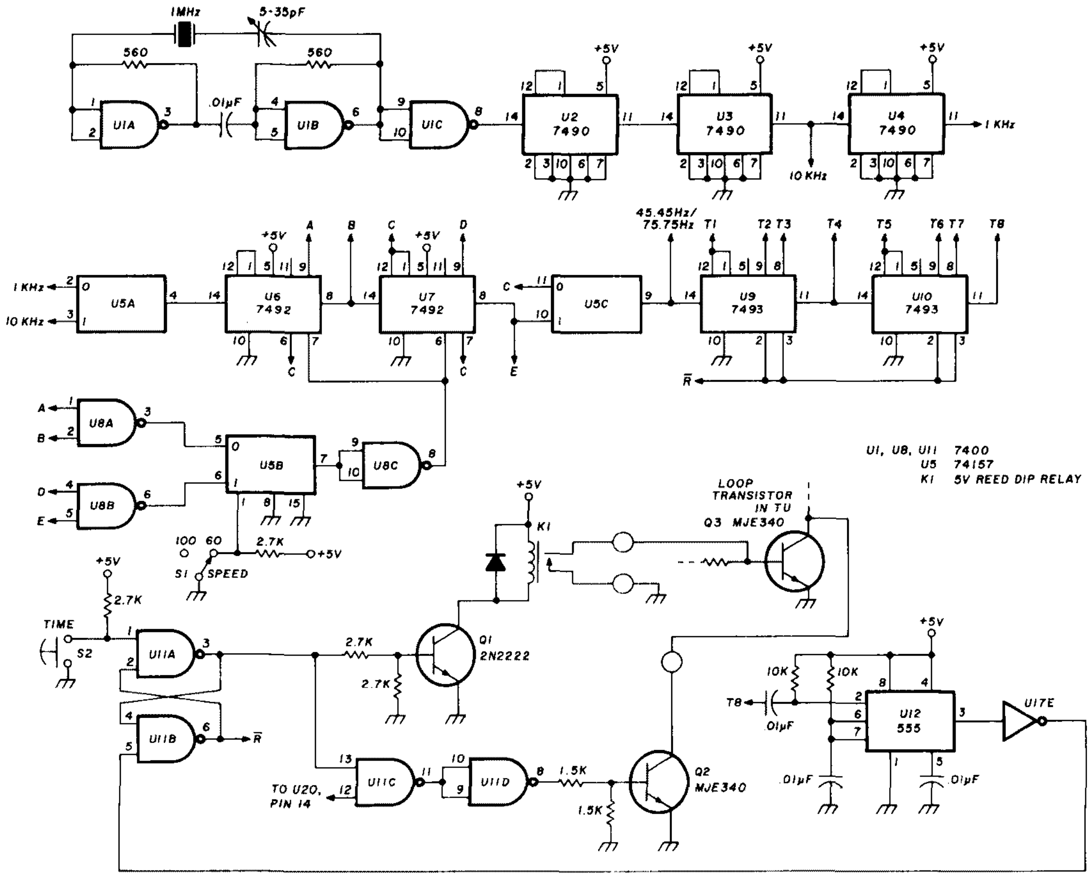


fig. 1. Timing and control schematic showing oscillator-divider chain, baudot divider, binary timing sequencer, and loop control circuits.

enables Q2, the loop transistor used during the time transmission. At the end of the time transmission T8 goes to zero, triggering one-shot U12. U12 resets flip-flop U11A, U11B, which ends the time transmission.

information transfer

U13-U16 (fig 3) are 16-position multiplexers that act as a 4-pole, 16-position switch whose position is controlled by timing pulses T4-T7. Table 1 shows how I connected the inputs from the clock and the hard-wired control characters. The multiplexers sequentially look at the BCD time from the clock, and the date from the thumbwheels and wired operations, then send this information through U18 to the PROM, U19, for baudot coding. Table 2 shows the relationship between the input address and the stored character conversions.

When T8 is low on the S input of U18, a 4-pole, 2-position multiplexer, the outputs look at the zero inputs, and timing pulses T4-T7 operate the PROM directly. T8 is also sent to the PROM (A4) and addresses the first half of the PROM. The first half of the PROM is sequentially addressed by T4-T7, reading out the programmed message from table 2. When T8 is high, U18 S input is high, and U18 switches to read the 4-bit information from multiplexers U13-U16. When high, T8 also switches the PROM to read the second half of the memory (table 2). The PROM is addressed randomly according to the input information.

The PROM has 8 output bits, but I used only 5 (the needed information bits). The other 3 bits could be used for the start and stop bits but are hard-wired in the output multiplexer, U20. U20, a 1-pole 8-position multiplexer controlled by T1-T3, is sequenced through at every address selected by the PROM and adds the start and 2 stop bits to the 5 information bits.

circuit operation

Set S1 to the machine speed. Depress S2. U11 goes high and turns on Q1, which energizes relay K, turning

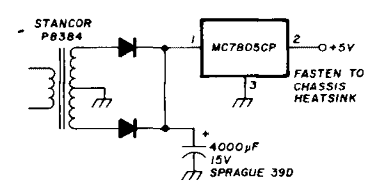


fig. 2. Regulated power supply schematic.

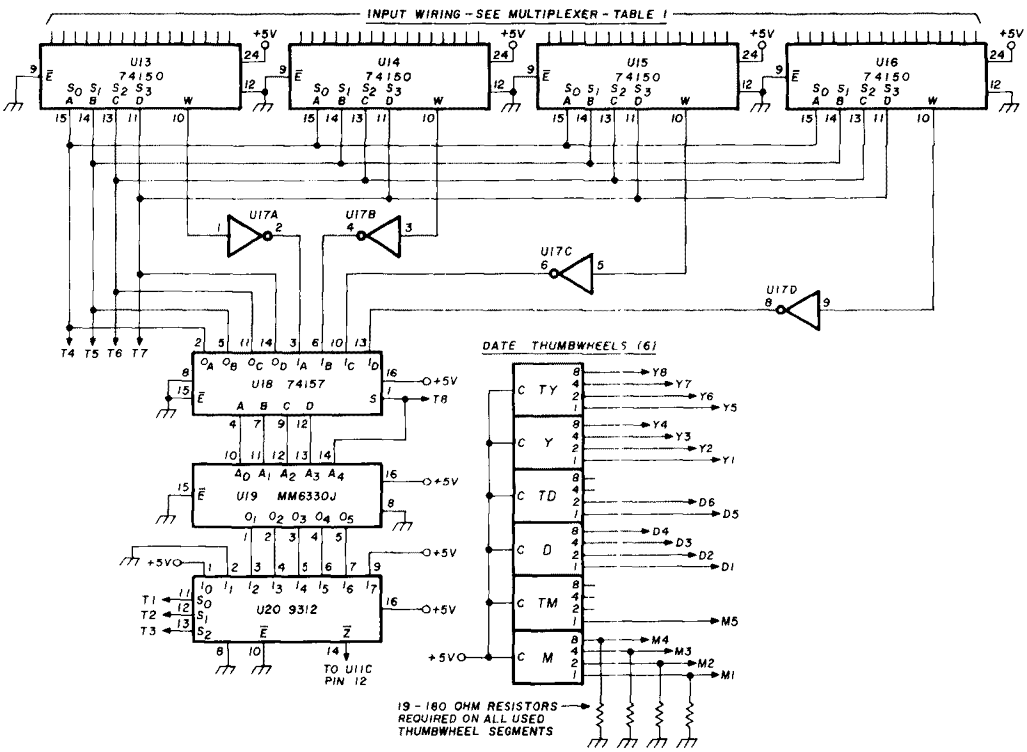


fig. 3. Information transfer schematic. Output is 32 8-bit words. The PROM outputs 16 words directly; the other 16 words may be programmed for the desired time format.

off loop transistor Q3. U11C is also enabled which turns on loop transistor Q2. U11B goes low, enabling U9, U10, which starts the timing sequence.

With T1-T8 at 00000000, the first character is ready. U18 reads T4-T7 and U19 reads the first half of the memory. T1-T3 sequence through U20, reading the first memory location (00000). The SPACE character (table 2) is serially output on U20 pin 14. When T3 returns to

table 1. Input connections from clock, and hardwired control characters.

9312 pin no.	74150 pin no.	address S3 S2 S1 SØ T7 T6 T5 T4	multiplexers U16	U15	U14	U13	comments
9A	1	0 1 1 1	GND	GND	GND	M5	tens, months
7A	2	0 1 1 0	+5V	+5V	GND	+5V	dash (-)
6A	3	0 1 0 1	t3	t2	t1	tØ	unit, minutes
5A	4	0 1 0 0	GND	GND	t5	t4	tens, minutes
4A	5	0 0 1 1	t9	t8	t7	t6	unit, hours
3A	6	0 0 1 0	GND	GND	t11	t1Ø	tens, hours
2A	7	0 0 0 1	+5V	GND	+5V	+5V	FIGURES
1A	8	0 0 0 0	+5V	GND	+5V	GND	SPACE
*	9	— — — —	GND	GND	GND	GND	ENABLE
15A,B	10	— — — —	U17A	U17B	U17C	U17D	output pin
------	11	— — — —	T7	T7	T7	T7	S3 input (D)
8A,B	12	— — — —	GND	GND	GND	GND	ground pin
13A,B	13	— — — —	T6	T6	T6	T6	S2 input (C)
12A,B	14	— — — —	T5	T5	T5	T5	S1 input (B)
11A,B	15	— — — —	T4	T4	T4	T4	SØ input (A)
9B	16	1 1 1 1	+5V	+5V	+5V	+5V	LETTERS
7B	17	1 1 1 0	Y4	Y3	Y2	Y1	unit, years
6B	18	1 1 0 1	Y8	Y7	Y6	Y5	tens, years
5B	19	1 1 0 0	+5V	+5V	+5V	GND	slash (/)
4B	20	1 0 1 1	D4	D3	D2	D1	unit, days
3B	21	1 0 1 0	GND	GND	D6	D5	tens, days
2B	22	1 0 0 1	+5V	+5V	+5V	GND	slash (/)
1B	23	1 0 0 0	M4	M3	M2	M1	unit, months
16A,B	24	— — — —	+5V	+5V	+5V	+5V	+5V pin

*not grounded on 9312s

0, T4 becomes a 1, addressing the second memory location (00001) that has the U character stored. This action continues for the words UNIVERSAL (SP) TIME until T8 becomes a 1, when U18 switches to read the time/date multiplexers, U13-U16, and U19 switches to its second half. U20, with T1-T3, reads the PROM memory locations addressed by U13-U16 as wired from **table 1.** T4-T7 now sequence the multiplexers through positions 0-15, reading the inputs as wired. At the end of the timing sequence (T1-T8 are a 1), T8 goes low starting one-shot U12, which resets flip-flop U11A, U11B, shutting off the time unit.

The date storage (**fig 3**) is on six thumbwheels. For those who would like to have the day changed automatically when the clock goes to 0000, the circuit of fig. 4A may be used. With this circuit S3 is changed monthly along with the month thumbwheel. ICs U13-U16 are 24-pin devices; and if 16-pin devices are desired, the circuit in **fig 4B** may be substituted, which uses eight

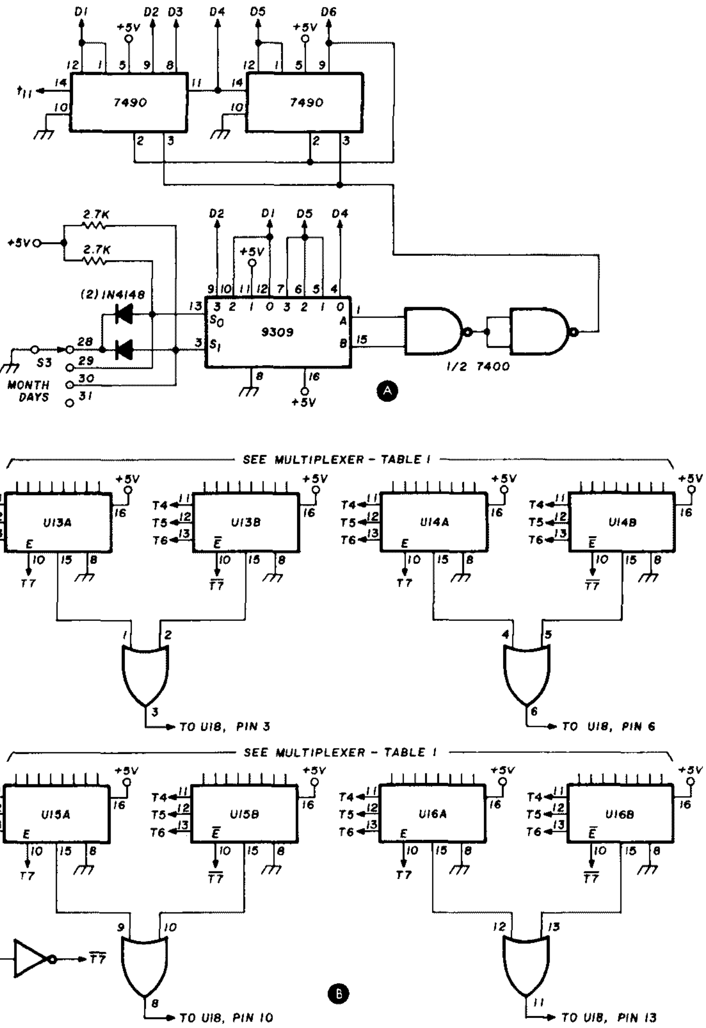


fig. 4. Optional circuits. Arrangement at A may be used for changing the day automatically when clock goes to zero; B shows how to use 16- instead of 24-pin multiplexers.

table 2. Conversions from input address to stored characters.

A4	A3	A2	A1	AØ	stored character	O_5	O_4	O_3	O_2	O_1	decimal address
0	0	0	0	0	SP	X	X		X	X	0
0	0	0	0	1	LT						1
0	0	0	1	0	U	X	X				2
0	0	0	1	1	N	X			X	X	3
0	0	1	0	0	I	X	X			X	4
0	0	1	0	1	V					X	5
0	0	1	1	0	E	X	X	X	X		6
0	0	1	1	1	R	X		X		X	7
0	1	0	0	0	S	X	X		X		8
0	1	0	0	1	A	X	X	X			9
0	1	0	1	0	L		X	X		X	10
0	1	0	1	1	SP	X	X		X	X	11
0	1	1	0	0	T		X	X	X	X	12
0	1	1	0	1	I	X	X			X	13
0	1	1	1	0	M				X	X	14
0	1	1	1	1	E	X	X	X	X		15
1	0	0	0	0	Ø		X			X	16
1	0	0	0	1	1		X				17
1	0	0	1	0	2		X	X			18
1	0	0	1	1	3	X	X	X	X		19
1	0	1	0	0	4	X		X		X	20
1	0	1	0	1	5		X	X	X	X	21
1	0	1	1	0	6		X		X		22
1	0	1	1	1	7	X	X				23
1	1	0	0	0	8	X	X			X	24
1	1	0	0	1	9			X	X	X	25
1	1	0	1	0	SP*	X	X		X	X	26
1	1	0	1	1	FG			X			27
1	1	1	0	0	:*	X				X	28
1	1	1	0	1	—	X	X	X			29
1	1	1	1	0	/				X		30
1	1	1	1	1	LT						31
T8	T7	T6	T5	T4							

PROM address (A4–AØ); output bit coding (O_5–O_1)

character input (addresses 0–11); time/date printout (addresses 20–27)

input timing

*not used

X=burn a "0"

8-position multiplexers in pairs with an OR gate. A regulated power supply is shown in **fig. 2.**

construction

A breadboard PC card is available that holds 36 chips for either 14/16- or 24-pin devices.* Both cards have IC foil patterns and the +5 volt and ground buses; also, each has a 44-pin connector (22 pins each side). The connector number is 44C.

Use number-26 AWG (0.3mm) solid wire and bypass each chip with a 0.05 μF disc capacitor and each row of 5-6 chips with a 47 μF tantalum capacitor. The finished card and power supply should be housed in an rf-tight enclosure. Bypass each side of the line where it enters the box with a 0.01 μF disc capacitor to chassis, and use shielded cable into and out of the box, or else rf can sneak in and foul up the logic — TTL makes a good detector!

ham radio

*Douglas Electronics, 718 Marina Boulevard, San Leandro, California 94577. Specify part number 11-DE-5 for 14/16-pin or 12-DE-5 for 24-pin devices.

fm detectors

The development of fm demodulators is traced from past to present — also included is a rundown on IC availability

The fm transceiver-repeater scene on the amateur vhf bands is an unusual part of amateur radio history. The technology was commercially developed and used by amateurs. This is in contrast with most other communications techniques in which the amateur contributed to development, and techniques were later applied to commercial use. The growth of amateur fm transceiver-repeater activity resulted from ingenious adaptation of cast-off 30-50, 150-170, and 450-470 MHz commercial equipment. The result is that today you'll find just about every kind of equipment in use from the oldest transceiver with loctal-based tubes to sophisticated designs using ICs, fets, and silicon power transistors. All have one thing in common: a circuit to convert fm signals to audio frequencies. In this article I've attempted to cover most of the practical fm detectors that have been developed. Some have been omitted but these have seen only limited use and remain as curiosities in the literature.

slope detection

Anyone who's tuned the two-meter band with an a-m receiver is aware that fm may be demodulated by an a-m detector. As you tune to an fm signal the demodulation is fairly effective until the receiver S-meter is at or near maximum; that is, the audio is clearest and loudest when you're tuned off to the side of the fm signal. This is called slope detection.

The fact that the a-m (diode) detector works at all on the fm signal is because of the receiver i-f characteristic. If the fm signal carrier is at point **A** or **C** on the i-f selectivity curve (fig. 1) frequency deviations will be converted to changes in amplitude. If the carrier is at point B, frequency deviations will not be converted to amplitude changes, especially if the i-f response is flat and symmetrical in its passband (as it should be for an a-m receiver).

Travis detector

A modification of the slope detector is the Travis detector (fig. 2). By using two resonant circuits offset by a small frequency difference, $f2\text{-}f1$, each driving a diode detector of opposite polarity, the sum of the diode detectors can give a near-linear output. **Fig. 2B** shows how the nonlinearities of the two (identical) resonant circuits cancel, in much the same way as in other push-pull circuits. The composite amplitude versus frequency curve of the Travis detector, when carefully adjusted, is much like that of the more familiar curves of the Foster-Seely discriminator or ratio detector.

Foster-Seely discriminator

The Foster-Seely discriminator is one of the most

By Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

common forms of fm detector, although in recent years the ratio detector has become more popular. The basic Foster-Seely discriminator is shown in **fig. 3**; variations are possible, as shown in **fig. 4**. The characteristic of the discriminator is the connection from the top of the tuned-transformer primary to a center tap of the tuned secondary circuit and the fact that the diodes are connected pointing in the same direction as in a full-wave rectifier. The explanation of discriminator operation is a bit tedious and usually involves phase vectors and "parallelogram of forces" diagrams.[1]

ratio detector

The ratio detector was developed later than the discriminator and is shown in **fig. 5**. The ratio detector is so called because it is sensitive to the ratio of the two voltages developed from the applied signal rather than to their difference. Since the ratio between the two voltages is the same at any level, an ideal ratio detector does not respond to a-m at any frequency deviation.

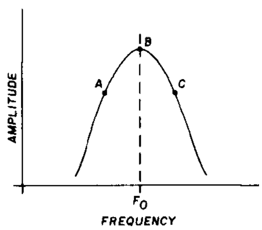


fig. 1. A-m receiver i-f response. An fm signal's audio may be recovered at point A or C on the selectivity curve.

The version shown in fig. 5A is the balanced circuit. Note that unlike the discriminator no direct rf connection is made between primary and secondary; rather a third or tertiary winding is used. Also note that the diodes point in opposite directions and that a large electrolytic capacitor is used in the diode load system to maintain a constant voltage during af variations. A variation of the ratio detector is shown in fig. 5B.

The ratio detector has one great advantage over the discriminator: it is relatively amplitude insensitive. Whereas discriminators are generally preceded by one or more stages of limiting amplifiers, the ratio detector is often used without any limiter stage ahead of it. The one awkard feature of the ratio detector is that the load capacitor must be disconnected during sweep frequency alignment. The detailed explanation of this detector's operation is similar (in approach) to that of the discriminator and is covered in the same chapter of reference 1.

injection-locked demodulator

The Bradley detector is one form of injection-locked oscillator fm demodulator.[2] A special tube (the Philco FM1000) was developed for this detector, and the circuit had a great deal to recommend it when these special pentagrid tubes were available. The Bradley detector is shown in fig. 6. Note that the cathode and first two grids of the FM1000 act as a Hartley oscillator in much the same fashion as many other electron-coupled oscillators. The Bradley detector requires no limiters ahead of it and

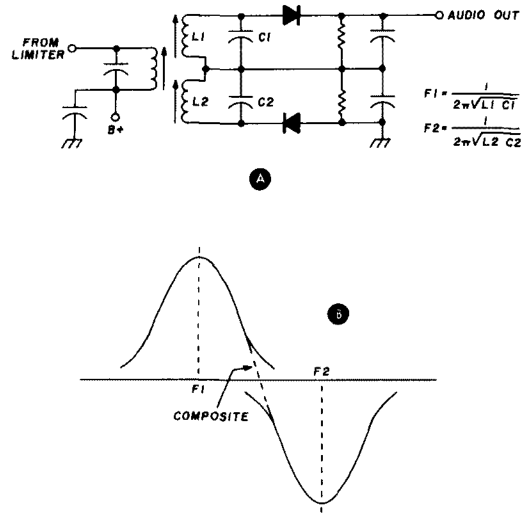


fig. 2. Circuit of the Travis detector (A) and its response (B) as a function of a frequency difference, $F_2 - F_1$.

has the additional advantage of very good output amplitude linearity versus frequency deviation. Note that the regions of amplitude beyond the linear portion (fig. 6B) are those where the oscillator is not being injection synchronized by the incoming signal (i.e., beyond lock-in range). The Bradley detector is one example of the injection-locked oscillator fm detectors described almost a decade earlier by Woodyard.[3] Woodyard did not have the special FM1000 tube available so he used a simple triode oscillator and diode detector as shown in **fig. 7**.

An injection-locked fm detector that achieved relatively large use in television sets as the sound detector was built around the 6DT6. This circuit is shown in **fig. 8** and is somewhat simpler than the Bradley detector, both in number of tube elements and number of resonant circuits. The oscillation mode is that of the tuned-plate tuned-grid, with the suppressor grid operating as a plate in the oscillator circuit.

gated-beam detector

This circuit was a popular fm detector, especially in television sets, before digital logic devices came into use. A gated-beam detector using the 6BN6 tube is shown in fig. 9. This tube was especially designed for this use and is a form of AND gate. Limiting is introduced because of symmetrical transconductance with respect to a fixed value of bias on the first grid. A tuned circuit develops a voltage whose phase is a function of the incoming signal

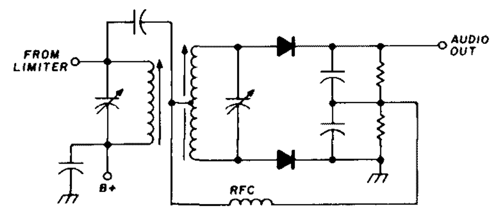


fig. 3. Basic Foster-Seely discriminator.

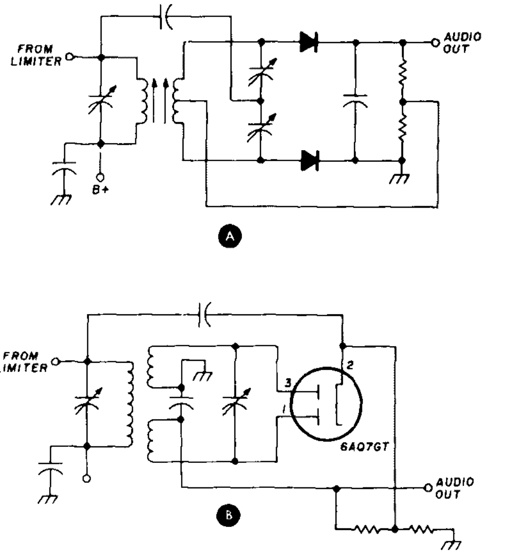


fig. 4. Variations of the Foster-Seely discriminator with capacitive secondary center tap (A) and dual-diode tube with common cathode (B).

frequency, producing discriminator action. An elaborate explanation of this circuit is given in reference 4.

pulse-counting detector

Another circuit that's seen some commercial use is the pulse-counting detector. Fig. 10 shows a pulse train, its differentiated components, and the resultant waveform (positive spikes) that are used to trigger a one-shot. The integral of the pulses is proportional to frequency deviation, since each pulse contributes a constant positive charge to an integrating capacitor. The voltage on the integrating capacitor is then propartional to the number

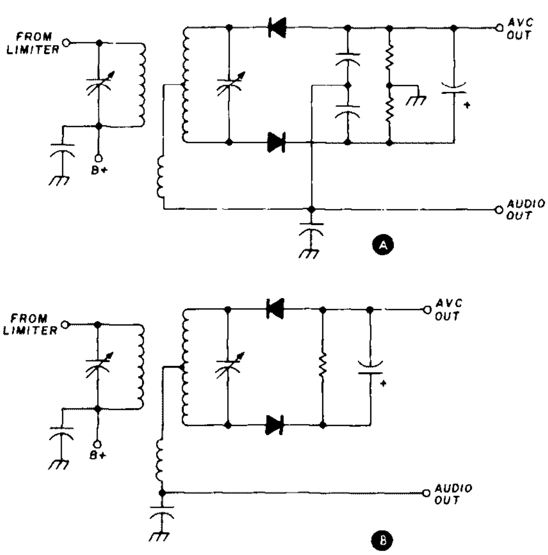


fig. 5. Ratio detector circuit, balanced form (A) and unbalanced form (B).

of pulses per second. The early versions used tubes and operated at an i-f of about 200 kHz.[5] A later version is described in reference 6. These low-frequency detectors are shown in fig. 11.

A pulse-counting detector operating at 10.7 MHz was developed in 1967.[7] The circuit used fast switching

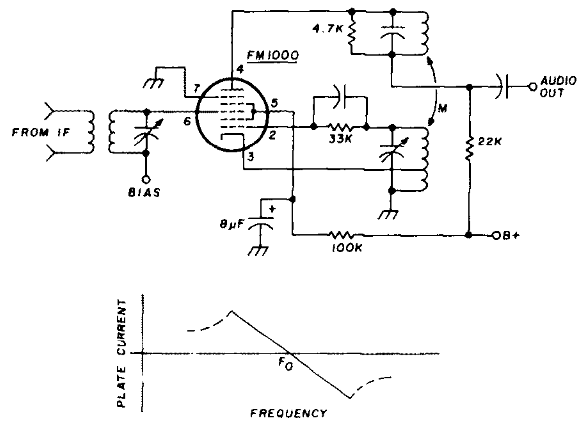


fig. 6. The Bradley detector and its frequency response.

transistors and a delay line to generate a short (25 ns) pulse (fig. 12). The delay line is used in much the same manner as in radar modulators to create a short pulse. A positive (limited) half cycle of the i-f waveform drives the delay line. CR1, CR2 conduct for a period that represents twice the delay-line length. At this point the delay line appears as an open circuit, since it's completely charged, and stops conducting. Q2's output, which is averaged by the 18k, 120 pF RC time constant, is proportional to frequency deviation.

IC fm demodulators

Digital IC one-shots have made their appearance as the pulse-generating portion of a pulse-counting fm detector. The IC one-shots perform well, but until recently the fastest devices have been the Fairchild 9601 and the Texas Instuments SN74121N and SN74122N. These one-shots have minimum pulse lengths of about 40 ns, barely fast enough for 10.7-MHz pulse-counting detectors. The technique of using TTL one-shots in fm detectors at lower frequencies is described in application

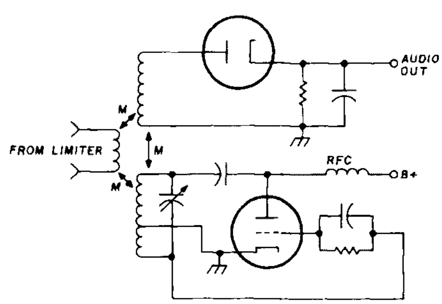


fig. 7. Injection-locked oscillator demonstrated by Woodyard in 1937.

notes.[8] Such a low-frequency pulse-counting fm detector is shown in **fig. 13.**

Note in this circuit how two other IC's (710 and 709 linear devices) are used as a limiter and averaging integrator respectively. Recently a second-generation Schottky

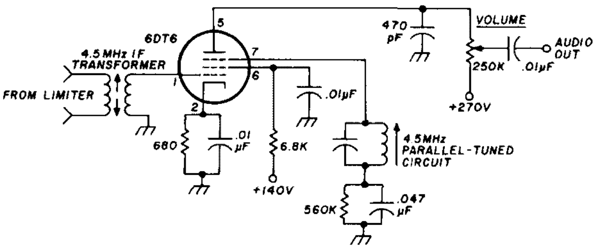


fig. 8. Injection-locked oscillator fm detector popular in early television sets.

clamped TTL dual one-shot, the 96S02, became available from Fairchild, which makes pulse-counting fm detectors much easier to design at 10.7 MHz. The minimum pulse length of the 96S02 is specified as 27 ns.

With IC one-shots in a pulse-counting detector, the IC Pandora's Box was opened. As mentioned in the section

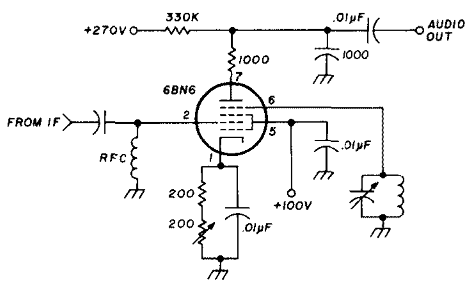


fig. 9. Another popular fm detector used in television sets, the gated-beam detector.

on the 6BN6, an AND gate can function in a quadrature-frequency modulation detector. Since a positive logic AND gate is functionally identical to a negative logic OR gate, either AND or OR gates will work. The operation is shown in **fig. 14** using square waves. Note

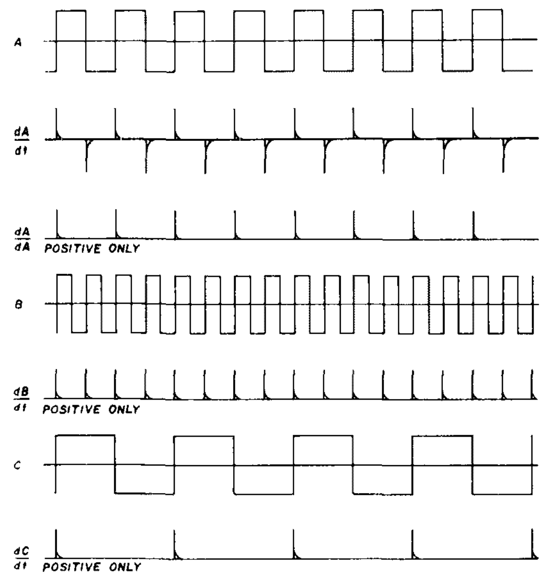


fig. 10. Waveforms in the pulse-counting detector. The positive-only diffferentiated spikes from three frequencies (A, B, C) give an output proportional to frequency.

that when **C** lags in phase (using the **B** quadrature as the reference), then the average AND gate output is increased. In opposite fashion, when D leads in phase from the reference position, the average AND gate output is decreased.

The μA717, μA718, and μA719 linear IC's by Fairchild are quadrature detectors using an OR gate. The μA717 and μA718 were introduced in **1967** with the first monolithic IC op amps (the μA702 and μA709). and so are probably the fm detector ICs deserving "granddaddy" titles. The μA717, μA718, and μA719 are shown in fm detector circuits in **fig. 15.** Note that, like the 6BN6 tube-type detector, only a simple parallel-tuned resonant circuit is required to provide quadrature. These Fairchild ICs were made as OEM sales components, so they rarely became part of experimental or amateur equipment. They are found in commercially made equipment, however.[9,10] Fairchild no longer lists the μA717, 718, or 719 in its current price sheets; however,two newer quadrature fm detector ICs are available: the

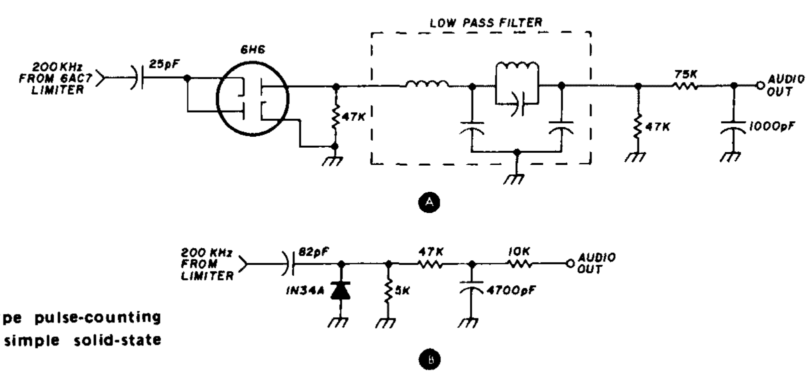
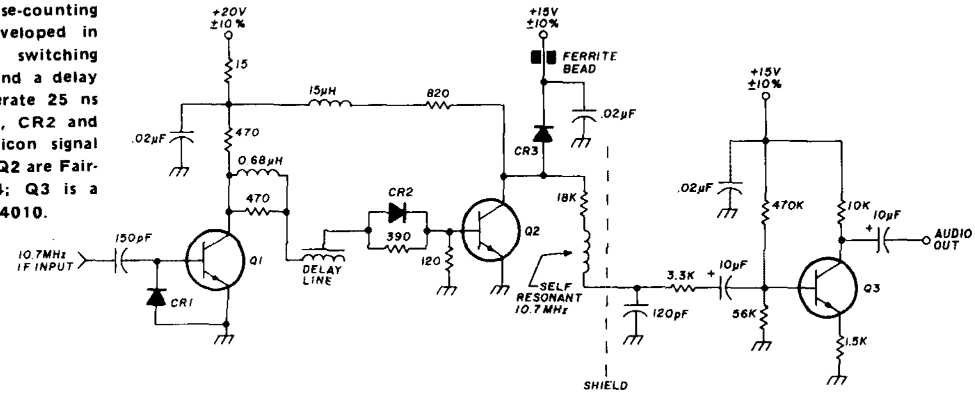


fig. 11. Early tube-type pulse-counting detector (A) and a simple solid-state version (B).

fig. 12. Pulse-counting detector developed in 1967 using switching transistors and a delay line to generate 25 ns pulses. CR1, CR2 and CR3 are silicon signal diodes; Q1, Q2 are Fairchild S1374; Q3 is a Fairchild SE4010.



μA754 and μA784. The latter is a second source for a European IC, the TAA640.

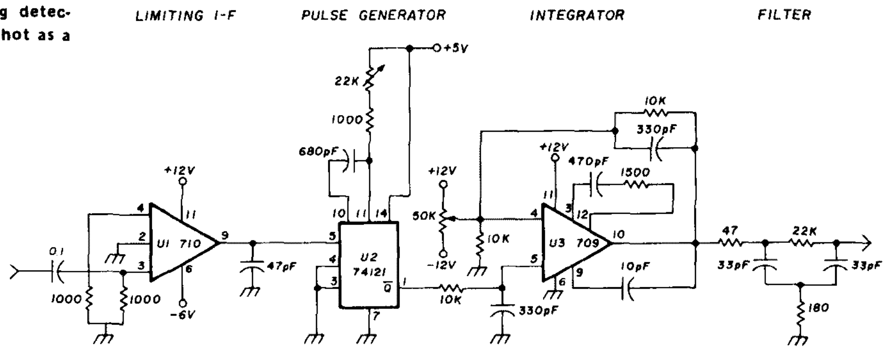
At about the same time Fairchild introduced their μA717 and μA718, Sprague described their version of the quadrature detector IC.[11] The ULN2111A by Sprague is perhaps the most widely used fm detector IC of all and is second sourced by at least five other IC manufacturers. The circuit using a ULN2111A is shown in fig. 16.

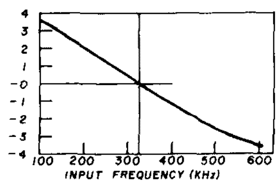
The fact that only a parallel resonant circuit is needed to obtain quadrature means that quadrature detectors such as the ULN2111A may be used at almost any frequency with a relatively simple frequency-determining element. I have seen the ULN2111A used at 40 kHz in an ultrasonic Doppler detector and even at audio frequencies in a tape-recorder flutter-measurement system. The ULN2111A has the following second sources: Fairchild μA2136; Motorola MC1357; National LM2111; Signetics ULN2111 (formerly N5111A) and RCA CA2111AE. Sprague has more recently released another more advanced IC called the ULN2113A, which is also a quadrature detector. The ULN2113A is also second sourced (as the LM2113A) by National Semiconductor.

Radio Corporation of America has had a line of linear ICs for fm detection for some years. The earlier ones were simply a combination of limiting i-f amplifier and diodes to implement a ratio detector or discriminator. These ICs (CA3013, CA3014, CA3041, CA3042, and CA3043) required the usual special transformers for these applications. The CA3043 is shown in fig. 17 as a ratio detector. Later ICs used the quadrature detector method; these are the CA3065, CA3075, CA3089, and CA3134. The CA3065 and CA3075 are second sourced by National Semiconductor as the LM3065 and LM3075. Motorola makes replacements for these two RCA units: the MC1358 and MC1375. Fairchild also makes a second source for the RCA CA3075, the μA3075.

Motorola has the MC1351 (second sourced by National Semiconductor as the LM1351) detector in addition to the second source types listed above, and National Semiconductor has the LM1808. National Semiconductor also has the LM373 and LM374, which are communications type ICs that will not only serve as fm detectors but also as a-m detectors or as product detectors for CW and ssb. The LM373 is shown in fig. 18 in its several configurations; note that in the fm detection mode it uses the quadrature detection method.

fig. 13. Pulse-counting detector using a TTL one-shot as a pulse generator.





phase-locked loop

ICs have been described that use the discriminator, ratio detector, and quadrature detector methods of fm

demodulation. It's natural to ask, "Didn't the locked oscillator ever get tucked into an IC?" The answer is yes and no. The injection-locked oscillator has all but disappeared in our technology and has been replaced by the phase-locked loop. Both types of synchronized oscillators may be used as fm detectors, but in IC technology only the phase-locked loop system has been used.

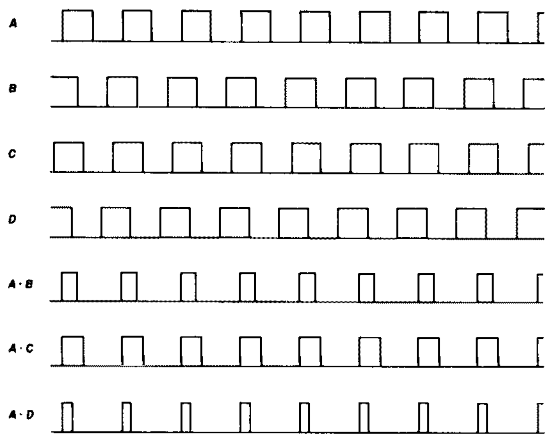


fig. 14. Representation of two square waves in quadrature in an AND gate. When pulse train C lags in phase with respect to B, average output is increased.

The phase-locked loop did not come into general use until the severe demands of space communications made it a necessity. The first reference to the phase-locked loop (in the U.S. open literature) was as a result of the Vanguard satellite program.[12] From its beginning in the mid-1950s, the phase-locked loop developed rapidly. Signetics introduced a one-chip version in the late 1960s.

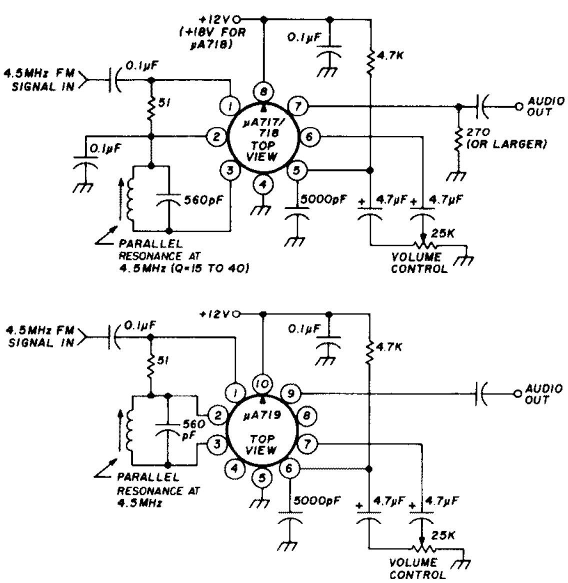


fig. 15. The Fairchild μA717/718 and μA719 as an fm detector and audio amplifier.

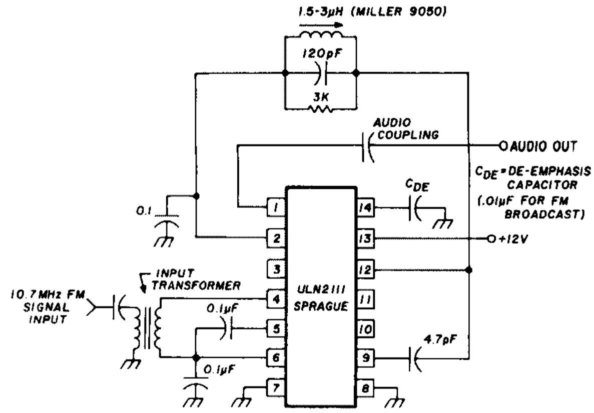


fig. 16. The Sprague ULN211 as a quadrature fm detector.

The Signetics NE560, NE561, NE562, and NE565 were monolithic IC phase-locked loops capable of a number of applications including fm detection.

The basic block diagram of the PLL is shown in fig. 19. Note that the loop locks to the carrier signal and deviations from this frequency are translated by the phase detector to deviations in dc average voltage. This phase detector output as applied to a low-pass filter whose output acts as a correction voltage to keep the voltage-controlled oscillator on frequency. The low-pass filter between the phase detector and vco is usually called the tracking filter and must be of special form if the phase-locked loop is to be stable.

The phase detector output is the audio recovered from the fm signal, assuming that the tracking-filter cutoff frequency will pass the audio frequencies of interest. An fm detector using the Signetics NE560 phase-locked loop IC is shown in fig. 20. Limiting ahead of the

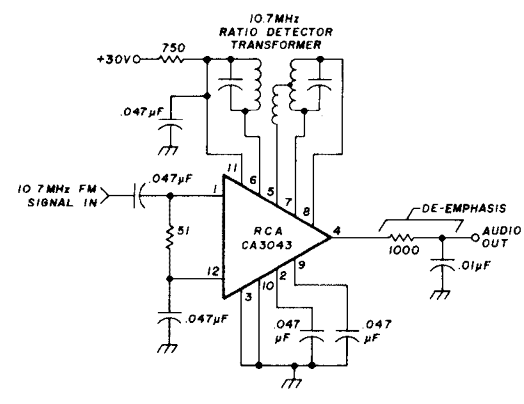


fig. 17. The RCA CA3043 IC as a ratio detector.

NE560 is provided in this case by another Signetics IC, the NE510. Note that the NE560 in fig. 20 has no inductors associated with it; the vco in this IC is an RC oscillator so no resonant circuits are required. This is true for the NE561, NE562, and NE565. The NE560, NE561, and NE562 will operate to at least 15 MHz,

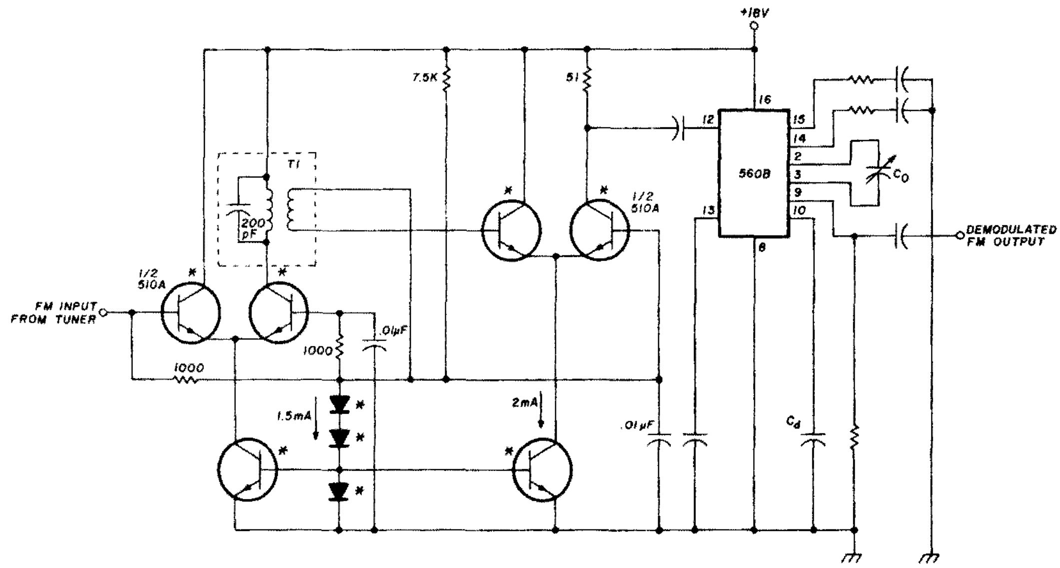


fig. 20. Fm detector using the Signetics NE560B phase-locked loop IC. Devices marked with an asterisk are part of a NE510A. T1 is bifilar wound on ½-watt, 100k resistor.

while the NE565 is only useful to 500 kHz. The NE565 is second sourced by National Semiconductor as the LM565.

Motorola also has a phase-locked loop IC family. The MC4024 and MC4044 comprise a dual vco and a phase detector respectively. Like the Signetics units, the vcos in the MC4024 are RC oscillators and only a capacitor is needed ot determine the oscillation frequency for a given value of vco control voltage. Since the MC4024-MC4044 pair is also used for frequency synthesizer circuits, it is TTL compatible and listed with Motorola's MC4000 series of TTL ICs. The MC4024 is

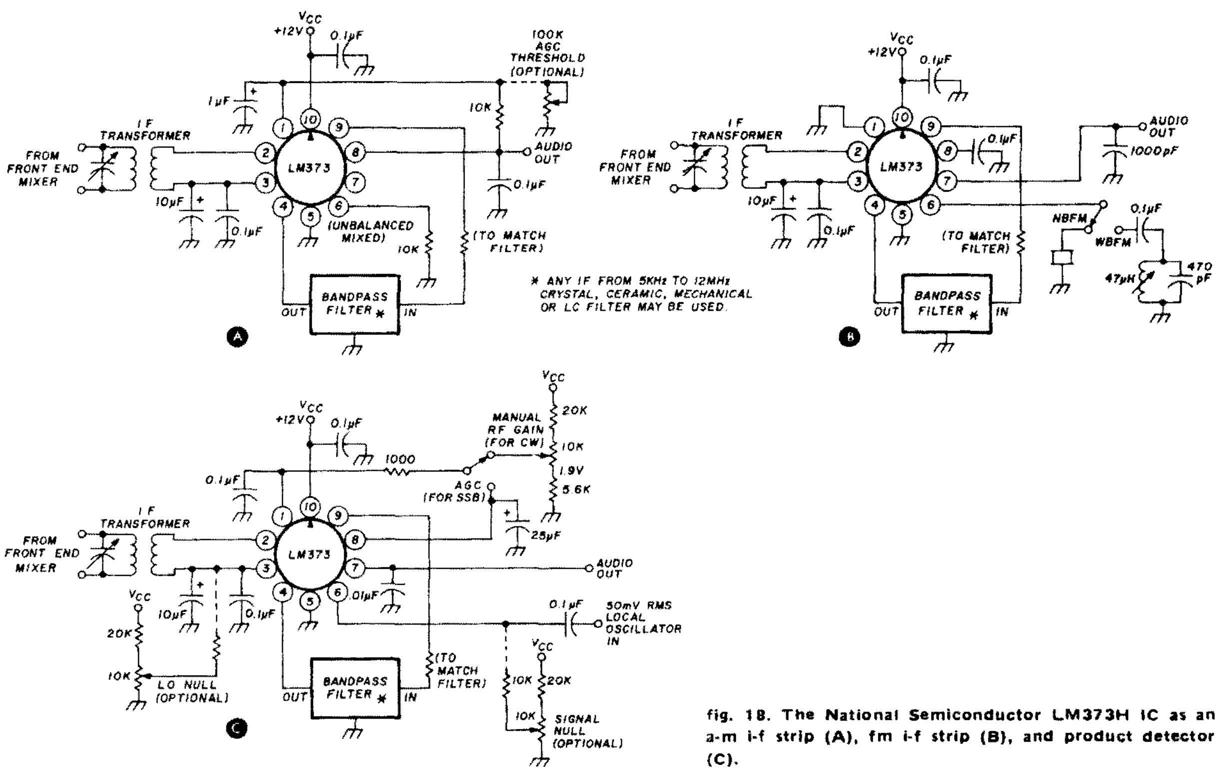


fig. 18. The National Semiconductor LM373H IC as an a-m i-f strip (A), fm i-f strip (B), and product detector (C).

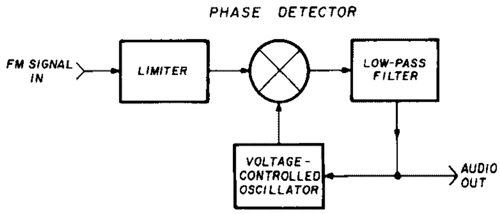


fig. 19. Phase-locked loop block diagram.

usable to 25 MHz. Fairchild has the μA780 phase-locked loop IC. This device is aimed principally at the color television market to be used to lock onto the 3.58-MHz color-burst frequency. The μA780 should work as an fm detector also, at least to 4.5 MHz.

The phase-locked loop ICs described above are designed for general PLL use or for some special purpose other than fm detection. Recently, Signetics has released the NE563, an IC PLL specifcally designed as an fm detector. In fact, the NE563 is designed as a complete i-f amplifier and demodulator for an fm receiver. The circuit is shown in fig. 21. Note that there is a down conversion from 10.7 MHz to 900 kHz, where the phase detector operates. I-f selectivity is provided at 10.7 MHz by a ceramic bandpass filter.

The phase-locked loop method of fm detection has some advantages over other methods in terms of signal-to-noise ratio. Even further increases in signal-to-noise ratio are possible using PLL techniques in more complex systems employing frequency compression. References 13 and 14 offer some light on signal-to-noise ratio advantages and frequency compression techniques.

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ham radio

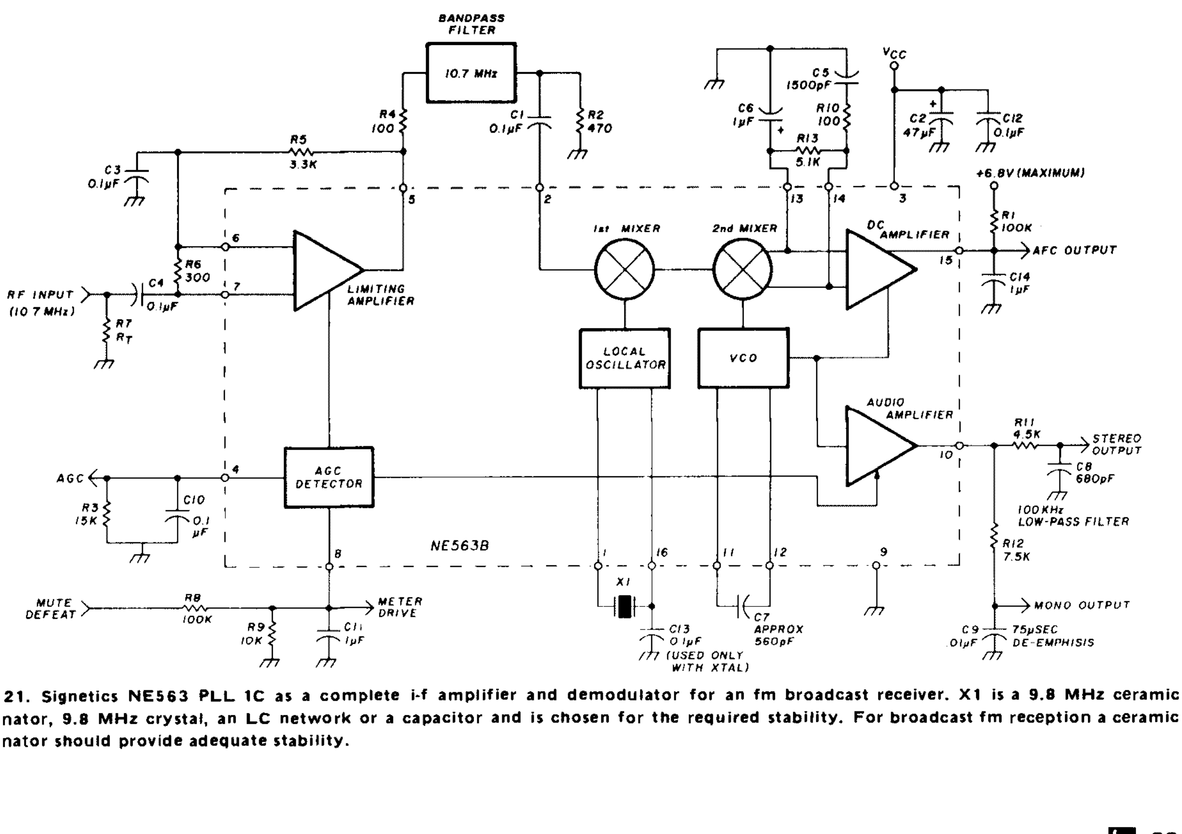
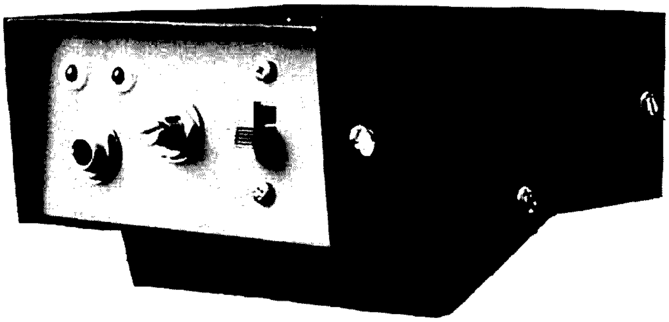


fig. 21. Signetics NE563 PLL IC as a complete i-f amplifier and demodulator for an fm broadcast receiver. X1 is a 9.8 MHz ceramic resonator, 9.8 MHz crystal, an LC network or a capacitor and is chosen for the required stability. For broadcast fm reception a ceramic resonator should provide adequate stability.



novel audio speech processing technique

offers maximum talk power with negligible distortion

A review of ssb speech processing techniques and a look at a new approach

Recently I was given the opportunity to preview a new speech processing unit which, to me at least, uses a different and novel design principle. I was impressed by the performance claim for the new unit, as well as by the incorporation of human engineering features which would enhance the performance and usefulness of speech processors based on other, older techniques.

The purpose of speech processing in transmitting equipment is to increase the average power output without exceeding the peak rating of the equipment. More specifically, the idea is to increase the amplitude of the weaker consonants in the human voice in relation to the louder vowel sounds. If this is done without introducing too much distortion, the intelligibility of the transmission *will be* greatly *enhanced* at the distant end during periods when propagation conditions are poor. This enhancement or increase in "talk power" can be as much as 10 to 12 dB according to recent literature (as well as articles dating back to the 1940s).

In recent years there has been renewed interest in devices which are essentially audio peak limiters, often known as speech clippers, logarithmic (quasi or otherwise) limiters, etc. This is not surprising when you consider the easy availability of inexpensive, miniature components. However, regardless of what these gadgets are called by their designers, they are essentially distorting devices, and this poses a problem in ssb systems. Several years ago, K1YZW took issue with such devices when they are used with ssb exciters;[1] I will try to summarize the points he presented.

In properly designed a-m and fm systems, speech or audio clipping (instantaneous limiting, logarithmic compression, or whatever) is useful as there is nothing to alter the critical phase relationship between the funda-

By Jim Fisk, W1DTY, Editor-in-Chief

mental signal and the harmonics generated in the process (see **fig. 1**). In ssb generators, however, this coherence, as it is generally called, is completely lost as can be seen in **fig. 2.** The fundamental signal will combine with the harmonics to produce peaks and valleys in the amplitude level. Using the numbers from Walter's article, these peaks can be as high as 1.7 times the original (limited) amplitude. To accommodate these peaks within the linear region of the transmitter, the exciter gain has to be reduced 5 dB, manually or automatically (alc), as compared to a sinusoidal input of the same peak amplitude! Although we want to increase our talk power, this means that we have to start with a 5 dB deficit – lowering our increased talk power by that amount to 6 dB or so.

Take a look at **fig. 3** which is a computer plot (submitted recently by K1YZW) of the rf envelope produced by a single clipped and filtered 400-Hz tone. This clearly shows the loww of power as compared to a pure sine-wave input. The latter, of course produces a CW signal with no envelope variations.

This amount of gain is still useful despite the very noticeable distortion in the modulation of the outgoing signal, although only little more than is obtainable with a well designed and virtually distortionless alc system. Unfortunately, in most cases a further reduction in exciter gain is necessary to accommodate the transients produced in the exciter ssb filter. K1YZW reports seeing an additional 4 dB overshoot from his Collins mechanical filter exciter with a square-wave audio input. Other exciters, especially those featuring filters with very steep skirts, may well be worse. With some exciters, most likely older ones, the problem may be less severe – it is obviously absent in phasing-type ssb generators.

I believe that the unknown and unpredictable transient behavior of ssb filters is the reason for the inconsistent results obtained with audio limiting; many amateurs find the process quite useless, while others using similar or even identical devices report a useful improvement in performance. If, on the average, we allow for total gain reduction of 7 or 8 dB, the few extra dB gain in talk power which remain hardly seem worth the effort since they are accompanied by the usual clipping distortion. As K1YZW and others have shown, a well designed alc system with properly chosen time constants will give a modest 2 to 3 dB of apparent talk power gain, free of distortion, and also provide a desirable safeguard against overmodulation or flat-topping.

It is worth noting that advocates of audio processors favor extremely severe cutting of the lower speech frequencies. Rolloffs of 12 dB or even 18 dB per octave, starting at 1.5 kHz, have been suggested.[2,3] K1YZW reported in a letter that in his experience with an essentially distortionless rf clipping system such drastic bass cutting has not proved desirable. In fact, he commented rather sarcastically that taking this approach to its ultimate limit will result in clipping only at the highest audio frequencies where it is not needed (and where it will produce virtually no power gain). It seems to me that severe bass cutting with its resultant lower talk power gain is used as a corrective measure against excessive low-frequency distortion, rather than to increase readability, though in final analysis these may amount to the same thing.

Clipping at rf rather than audio is free from these deficiencies. While harmonics are produced by the rf clipping process, they are far from the passband of interest so are easily removed by filtering, and maximum talk power is obtained without difficulty. Many articles have appeared on the subject of rf speech processing which should be referred to for more detailed information (see list of references at end of article). At least two manufacturers in the United States provide kits for modifying the more popular commercial exciters.

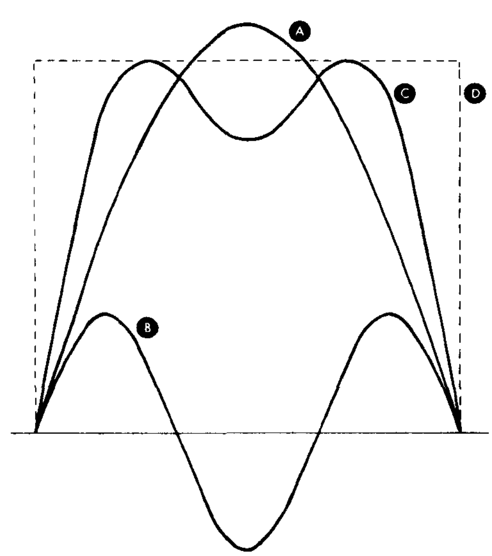


fig. 1. The derivation of a square wave. The fundamental component is shown in A. B is the third harmonic. The sum of A and B is shown in C. The square wave at D is the resultant with higher odd-order harmonics of proper amplitude and phase.

Clipping a double-sideband signal has been suggested occasionally,[2] but a little imagination will show that this is really no different from audio clipping. In a letter to me concerning ZL1BN's article, K1YZW mentions that, in his original work with rf clipping in 1961, he actually investigated this method using a high resolution spectrum analyzer to confirm this point. He also suggested that the good results which ZL1BN obtained with his circuit (**fig. 7**, page 33, reference 2) was due to the fact that it was not DSB clipping at all. His conjecture is based on the fact that most ssb filters show some selectivity at their input terminals where ZL1BN placed his clipping diodes. Unfortunately, you cannot rely on being able to duplicate ZL1BN's results because they seem to depend on a characteristic of the ssb filter which is not controlled by the manufacturer.

An interesting variation of rf clipping is the approach first used by Comdel, Inc., and described by me in 1968.[4] This is essentially a closed-loop ssb system with rf clipping so the output is at the original audio frequency. The advantage of this system is that it forms a self-contained unit and avoids the need to modify exist-

ing ssb exciters. The same principle is now being used by manufacturers in England and Japan.

Until recently I was firmly convinced that rf clipping was the only way to achieve a significant, distortionless increase in the talk power of an ssb transmitter. How-

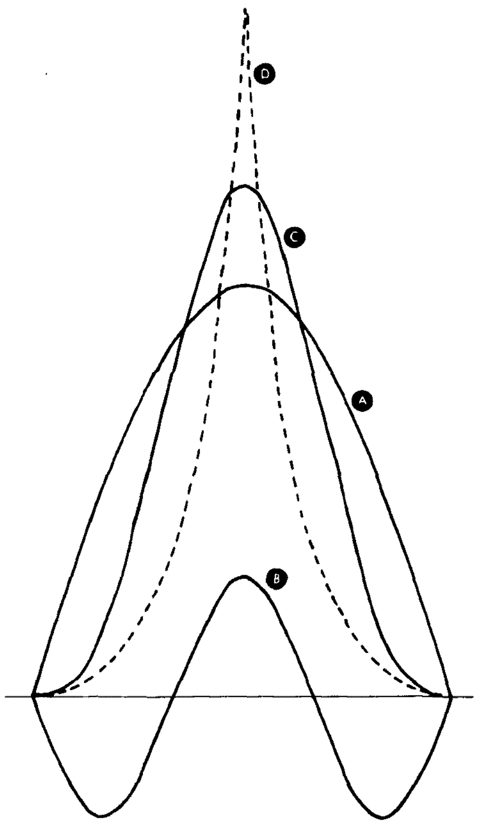


fig. 2. The effect of loss of phase coherence. A is the fundamental; B is the third harmonic; C is the sum of A and B; D is the resultant of all harmonics in the passband.

ever, this conviction was severely shaken by the presentation of a radically different approach to the problem in a unit produced by Maximilian Associates, a small consulting company here in New England.* After reading the specifications of the claimed performance, I found it hard to accept that I was looking at an audio processor. Following are the important operating features of the Maximilian Associates model SBP-3, as it is tentatively called:

1. Provides 16 dB of instantaneous limiting.

2. Harmonic distortion does not exceed 9%, and is typically around 5%.

3. Frequency response is within 1 dB between 460 and 2300 Hz (-6 dB points at 400 and 2600 Hz).

*Maximilian Associates, Box 223, Swampscott, Massachusetts 01907.

4. Shows sensitivity at limiting threshold of 2 mV rms with a signal-to-noise ratio of more than 30 dB.

5. Cannot be overloaded with inputs up to 1 volt rms.

6. Incorporates audio agc so that 16 dB of compression cannot be exceeded.

7. Features two front panel lamp indicators, one (green) signifying adequate input, the other (red) denoting an excessive gain setting.

The design philosophy behind this new speech processing device was described to me as follows: In K1YZW's paper[1] the conclusion was reached that any process which causes appreciable distortion of the audio signal is not suitable for application to ssb systems. The obvious question at this point is how large is "appreciable" or how much distortion can, in fact, be tolerated. If we accept an initial gain reduction of 1 dB to accommodate the amplitude peaks caused by the non-coherent addition of the distortion products, then the distortion of the *process* must be held below 10 per cent (note that straight audio clipping produces more than 40 per cent distortion). Such low distortion can hardly be perceived by the human ear in a speech communications set-up and is unlikely to cause any appreciable transients in the ssb filter.

In order to achieve such low distortion, the speech frequency spectrum is split up into four bands and each band is independently processed, i.e., peak limited and filtered. **Fig. 4** shows a block diagram of the scheme and the center frequencies of each band. Since each band is quite narrow, the distortion products fall (theoretically) outside its limits and can be removed by filtering. The

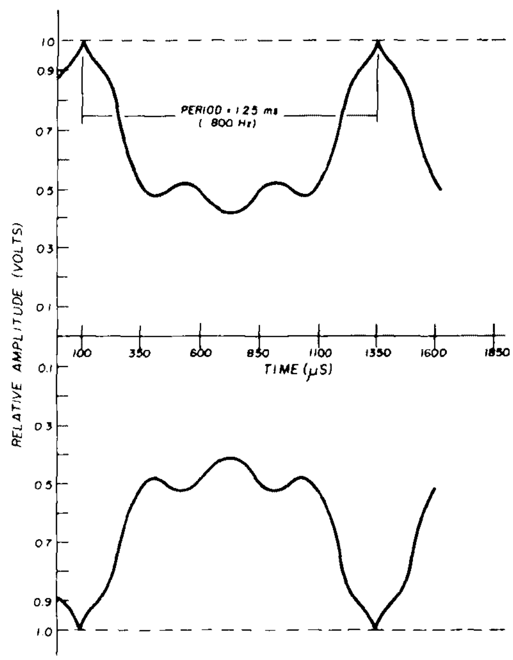


fig. 3. Envelope of rf signal with 400-Hz clipped and filtered input. Average power is -5.3 dB. Dotted line denotes envelope with sine-wave input; average power is 0 dB.

concept is straightforward and has been around for quite some time. As a footnote, it is probably fortunate that the designers did not discover some earlier references until after the first model of the SBP-3 had been successfully tested — H. Schneider[7] discussed this method in considerable *detail* 20 years *ago* and concluded that a working system was not economically feasible. This view seems to be shared by ZL1BN (though he does not even consider the major problem of the combiner stage which will be mentioned later).

At first thought, it would seem desirable that the bandpass filters should have steep skirts so that there would be practically no overlap between adjacent bands. This approach is extremely costly and, as it turns out, technically unsatisfactory. The post clipper filters would show the typical "ringing" effect when subjected to a near square wave input, and undo most of the clipping action as well as introduce considerable distortion. To avoid this a filter is needed with a Gaussian characteristic which, unfortunately, exhibits a rather gradual cut-off response. This type of response is approximated fairly closely by a low-Q tuned circuit (this is what was used in the new speech processor). The Q values and center frequencies of the individual bandpass filters were chosen to give a reasonably flat response between 400 and 2400 Hz. Of course, a considerable overlap exists between adjacent bands with this approach.

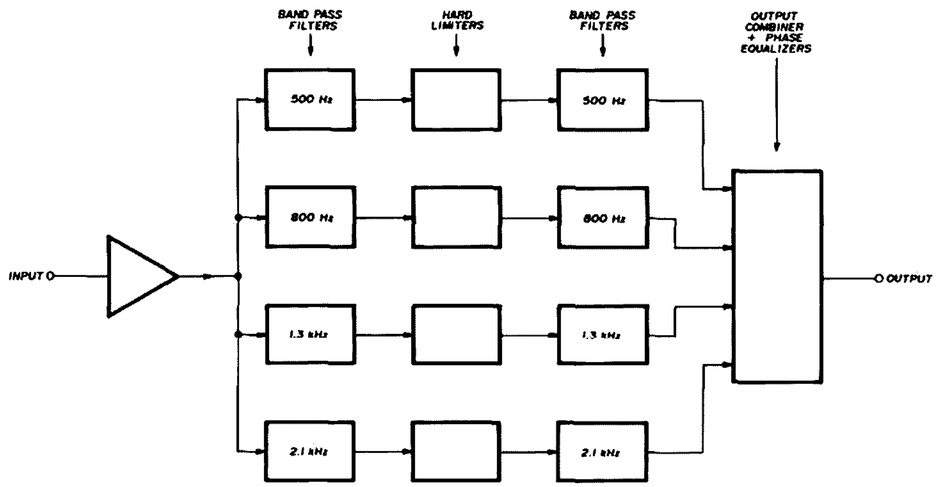


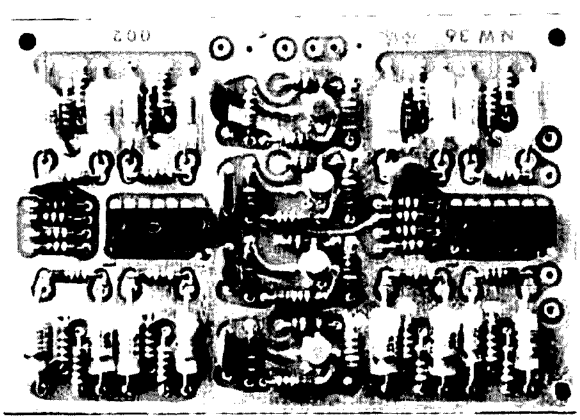
fig. 4. Simplified block diagram of the band-splitting speech processor. This processing scheme, as embodied in the Maximilian Associates SBP-3, provides 16 dB of instantaneous limiting with harmonic distortion not exceeding 9% (typically 5%). Frequencies are the center frequencies of the bandpass filters.

Contrary to initial expectations, the overlapping filter bands have proven quite satisfactory. A signal falling near the center of a band will have its clipping harmonics attenuated by more than 20 dB which is more than adequate. Signals which fall near the middle of the overlap region (i.e., half way between adjacent filters) will be clipped less severely and therefore less filtering is needed. The mathematics showing this are quite tedious, but clearly indicate that the distortion can be held under 10 per cent for any signal between 400 and 2400 Hz, relative to the limited output level. Great care was taken in the design of the limiters to obtain *near perfect symmetry.* This assures that only odd order harmonics are produced; the simple bandpass filters are not adequate to deal with any appreciable second harmonic content.

A final difficulty arises in the design of the combiner unit. A particular tone may easily produce near-equal outputs from adjacent bands. If these outputs were simply added together, partial amplitude cancellation may result due to the phase difference between them. To avoid this and to maintain a flat frequency response, the output combiner contains phase shifting networks.

Almost as impressive as the design of the processing circuitry are the ancillary features and attention to detail which makes the SBP-3 a truly deluxe instrument. The

Miniature circuit board for the Maximilian Associates SBP-3 includes four active bandpass filters and a combining circuit.



audio agc circuit at the input is virtually free of distortion up to one volt input and is so effective that 16 or 17 dB of compression cannot be exceeded. This makes the device essentially overload proof. Front panel indicators, in the form of LEDs, show that full compression has been reached (green) or that excessive gain is being used (red). The latter condition, if allowed to exist, is simply a warning that there may be considerable background noise during long speech pauses.

The threshold levels for the green and red indicators are 10 and 100 mV respectively; the output signal level of the SBP-3 is approximately 80 mV. Remarkably, there are no critical components employed in the circuits, the closest tolerance being 5 per cent, and there are no internal adjustments. This makes the device suitable for marketing in kit form and I believe that this is being considered by the manufacturer.

A not uncommon difficulty with solid-state transmitting accessories is rf feedback or local RFI. Curing this is often laborious and time consuming. In the SBP-3 this possibility has been virtually eliminated through the use of rf layout, construction and bypassing techniques. As an extra precaution, the unit is housed in a metal enclosure arranged for good additional shielding. The power requirements are modest, nominally +12 volts dc at 35 mA (voltages between 10 and 18 volts are acceptable and don't affect performance in any way).

How does the SBP-3 stack up against a well designed rf clipping system? Tests have been going on for over six months and results seem to favor the new device by a small margin. This is believed to be due to the agc circuitry in the SBP-3 rather than the difference in the processing techniques. It is worth noting that there is a subtle difference in performance which, depending on circumstances, may make either method slightly preferable. The band splitting approach in the SBP-3 is likely to produce less intermodulation distortion (beats between two co-existing, non-harmonically related tones) than rf processing methods; on the other hand, rf processing (in theory) is completely free from harmonic distortion which is present, though small, in the new approach. The SBP-3, as presented to me is a high quality, technically superior – though not inexpensive – accessory for the transmitting radio amateur.

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ham radio

SB-102 modifications

Too many CW signals sounded less than T9 on my Heath SB-102. Hum modulation in the receiver section was suspected but the audio section was clean and the CW monitor oscillator tone was pure. The trouble was traced to the wiring of the switched ac for the power supply. Heath uses unshielded wire laced into the cable to run from the power supply plug to the on/off section of the *function* switch on the front panel. This induced hum into the other wires in the harness. Also, it was noted the proximity of the wires at the rear of the switch to the envelope of V3 had a bearing on the amount of induced hum.

Two methods may be used to solve this problem. The first is to use shielded wire for the run between the plug and the switch section. Small diameter coaxial-type wire was used here. The second method is to disconnect the wires from the power plug pins and insert a jumper. The center-off switch on the HP-23B power supply is then used for turning the SB-102 on and off.

The Heath SB-102 provides one extra relay point for use with linear amplifier switching If more relay-switched functions are desired, an extra relay must be employed. A Potter & Brumfield KHP17D12VDC 12-volt dc relay was used in the circuit of **fig. 1.** These are similar in construction to the ones originally used in the SB-102. The 12 volt ac line is tapped at board 85-130, rectified and filtered by D1 and C1. Relay points 10, 6, and 2 of RL1 are used to switch the ground end of RL3. RL1-6 is removed from pin 11 of the power plug and similar points of relay RL3 are used to provide that switching function.

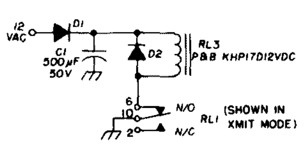


fig. 1. Circuit for adding a control relay to the Heath SB-102 for additional relay-switched functions (one auxiliary set of contacts is available in the original circuit).

The relay is socket-mounted on a small L-shaped bracket of aluminum fastened to the wall of the final amplifier rf cage. There is adequate room here for mounting the relay and routing the wires to and from the relay points.

When cleaning relays of the type used in the SB-102, it is best to remove the armature by unhooking the spring and gently prying back on the flat spring at the armature center point. Burned relay points not noticeable with the armature intact are now readily seen and cleaning is simplified. Use a burnishing blade designed specifically for cleaning relay points. Do not use excessive pressure on the bodies of these relays because the frame is made of soft metal and is easily bent. Alignment of the normally-open, normally-closed and armature points of the relay should be checked. See that the centers of the points contact each other. Gentle bending of the relay frame will align points that are off-center. Also, check to see if there is a small amount of "rise" to the armature points on the make to insure a slight wiping action. These relays should be checked periodically, especially if receive/transmit intermittents are encountered.

Paul K. Pagel, K1KXA

improved selectivity for Collins S-line receivers

A simple no-holes modification provides the equivalent of a 14-pole filter with a measured 1.54 shape factor

With the inevitably increased crowding of the high-frequency amateur bands combined with stronger signals from better equipment, receiver manufacturers have been acting to satisfy the need for increased selectivity through the use of better filters. Presented here is a simple means of increasing the voice-range selectivity of the S-line receiver to exceed that of most receivers on the amateur market.

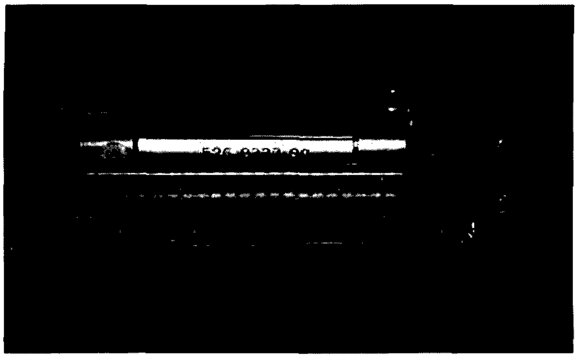
The cascade configuration of filters is a well-known method of increasing selectivity, which should theoretically improve skirt selectivity by doubling the attenuation. For proper results, it's necessary to employ a stage of isolation between the filters. In my case, this also serves to make up for the insertion loss of an additional filter.

approach

In my receiver, the S-line 75S3B, this concept becomes a relatively simple procedure. A second filter was added in the i-f strip as shown in **fig. 1**. The i-f gain control, R57, is used to touch up the calibration for the S-meter after the modification. Point A is where the signal was inserted for my measurements, and B is the output point. **Fig. 2** provides the circuit details regarding the additional filter, FL1, and its associated amplifier. The new filter is always operational, which precludes some problems and simplifies the modification.

Fig. 3A is the stock 2.1-kHz i-f passband of the S-line as a result of the Collins 455FA21 mechanical filter. This filter is a seven-disc unit, which is the equivalent of a seven-pole filter. **Fig. 3B** is the resulting improvement through the addition of a second filter, which was a 455Y21, the hermetically sealed version of the 455FA21. **Fig. 3C** is a double exposure of the two, superimposed for ease of comparison, which dramatizes

By Marv Gonsior, W6FR, 418 El Adobe Place, Fullerton, California 92635



Added filter and amplifier mounted on Vector board clad on one side. Isolation between input and output leads and existing filter below is important.

the improvement. The results of cascading the 1.5-kHz filter through the 2.1-kHz filter are depicted in fig. **3D.** It may be seen, as expected, that a modest improvement is obtained. **Fig. 3E** is the passband of the 0.5-kHz CW filter before matching by proper adjustment of the resonating capacitors, and **fig. 3F** is the very worthwhile result of proper termination. Peaking by use of the S meter is not adequate for proper resonating/terminating

which could be used instead of the more sophisticated equipment I used.

supply voltage

Regarding a source of V_{DD} for the fet, I used a 9-volt, 1-watt zener from the 72-volt screen supply for the i-f strip. This voltage was dropped through two 15k,

table 1. Performance comparison of the two filter systems.

mode	6 dB	60 dB	shape factor
one filter	2.12 kHz	4.70 kHz	2.22:1
two filters	1.97 kHz	3.03 kHz	1.54:1

1-watt resistors in parallel. The current requirement of only 4 mA is nominal. This may be met in numerous ways, such as dropping down from the 140-volt zener, CR6. The power consumption is the same. Just be sure that the V_{DD} supply is well-filtered dc and is adequately decoupled. One change I incorporated, as a result of seeing the 72-volt i-f screen supply move sharply upward as much as 20 volts coincident with agc voltage, was to regulate it with a 75-volt zener that is barely regulating but limits the upward excursion of this supply voltage. A 0.1 μF bypass capacitor across the zener eliminated the small noise it generated. The importance of this change

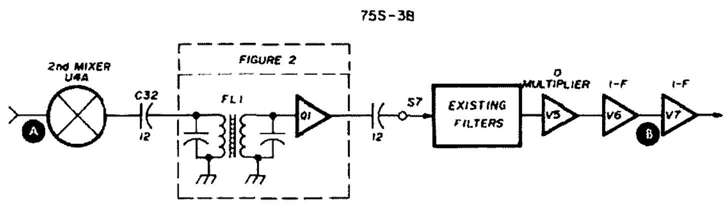


fig. 1. Collins 75S-3B i-f strip showing addition of the 455Y21 filter. Test signal for measurements was inserted at point A; output was taken at point B.

of the filter. Collins advises that it is quite important to resonate their filters at the *lowest* signal level possible to preclude saturation. **Table 1** summarizes the measured relative performance of the two systems.

measurements

My instrumentation included Hewlett-Packard equipment consisting of an 8443A tracking generator in conjunction with their spectrum analyzer made up of the 141 display and the 8553B rf and 8552B i-f units. An H-P 1121 fet probe was used to sample the signals. For ease of photography, the signal voltages from the spectrum analyzer were transferred to a Tektronix 7704 oscilloscope. All instruments were laboratory grade and have accuracies traceable to the National Bureau of Standards. The calibration scales on all scope pictures are 1 kHz and 10 dB per division. Measurements down to the noise at -75 dB were easily accomplished.

If you wish to duplicate my measurements, it's necessary to remove V8, the bfo tube, otherwise it will produce a broad signal on one or the other skirt; i.e., the upper or lower sideband. A very interesting and applicable article by Arthur Lux[1] describes a relatively simple system for observing filter bandpass characteristics,

is to correct the regulation, which tends to drive up the G_m of the tube while the agc is simultaneously trying to lower it! The same problem existed in the 32S1 exciter with its alc voltage.[2]

construction

The modification is easily accomplished without any drilling and results in the equivalent of a 14-pole filter

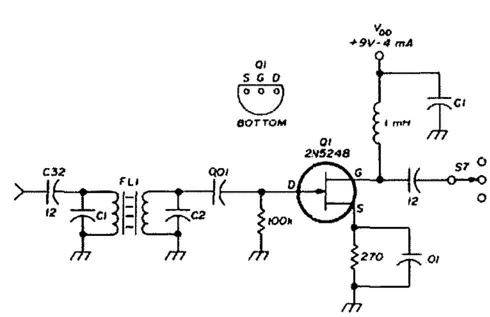
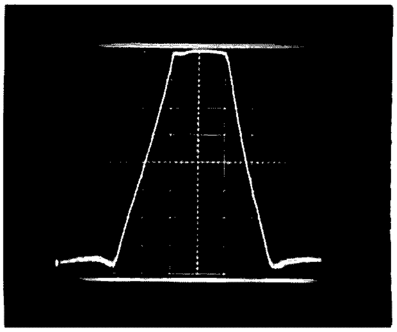
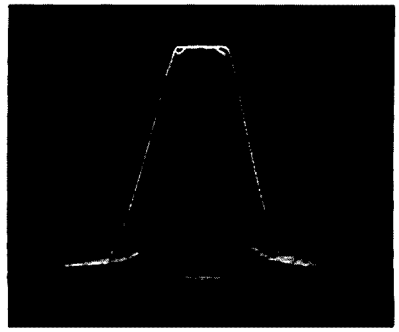


fig. 2. Circuit of added filter and its amplifier. V_{DD} for the fet was obtained through a 1-watt zener diode from the i-f strip 72-volt screen supply, which was decreased through two 15k, 1-watt resistors in parallel. C1, C2 are 130 pF, 1%, or trimmers.

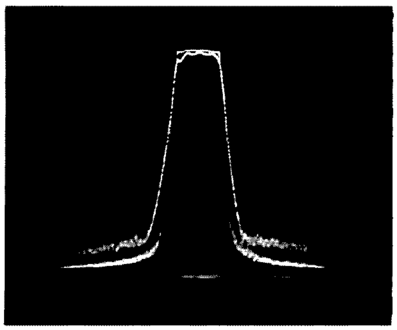


A. 2.1-kHz i-f passband of the stock 455A21 mechanical filter.

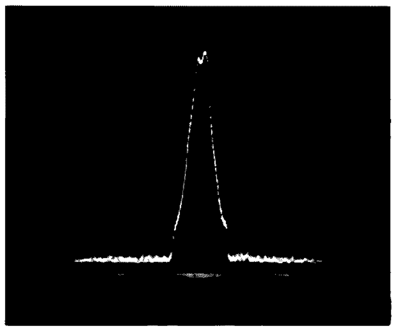
B. Passband improvement by adding the 455Y21 filter.



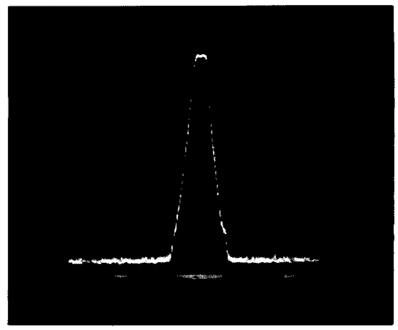
C. Traces A and B superimposed for comparison.



D. Results obtained from cascading the 1.5-kHz filter through the 2.1-kHz filter.



E. 500-Hz CW filter response before proper termination.



F. Passband response with filters correctly resonated and terminated.

fig. 3. Oscilloscope traces obtained during measurements, which show improvements obtained by adding a second filter. Vertical and horizontal scale calibrations are respectively 10 dB and 1 kHz per division.

with a 1.54:1 shape factor. The basic cost of the change is that of the additional Collins mechanical filter. The 455FA21 (part no. 5269427-000) currently retails for $59.88. The equivalent filter in a hermetically sealed case is the 455Y21 (part no. 5269337-000), which retails for $96.13. Both units may be found in surplus and other markets at relatively favorable prices.

I constructed my unit using a pad cutter on a piece of Vector board clad on one side. It measures 7/8 x 3-1/8 inches (22x80mm). Layout isn't critical as long as certain basic rules are followed. These are: keep the input and output leads apart, remembering that you're working with −75 dB signal levels; use a ground plane to shield the new filter from the ones below it; and keep the drive levels low to avoid overdriving the second filter. The total job should take about two hours. The board mounts easily on its own leads and with a lug to a ground terminal, which conveniently protrudes inside the shield around the bottom of T4 and the existing mechanical filters (fig. 4).

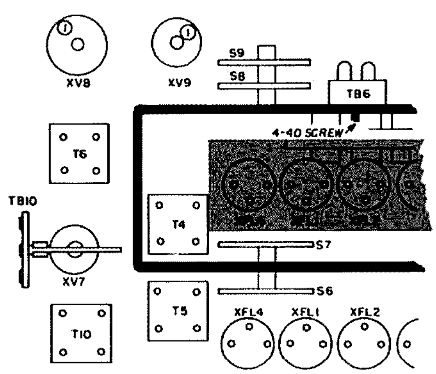


fig. 4. Shaded area shows location of filter circuit board on the S-line receiver. Board is secured to a solder lug attached to TB6 ground terminal.

results

The benefits of using the two filters in cascade, properly terminated, are quite impressive as the scope pictures show. For example, the frequency excursion required for my S9 +40 dB crystal calibrator signal to become inaudible was reduced by 750 Hz with two filters installed versus one. In digging out weak signals in interference, this filter system is quite effective, as the theory predicted.

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2. P. Kreager, W6UYW, and M. Gonsior, W6VFR, "32S1 Operating Improvements," *CQ*, August, 1962, page 47.

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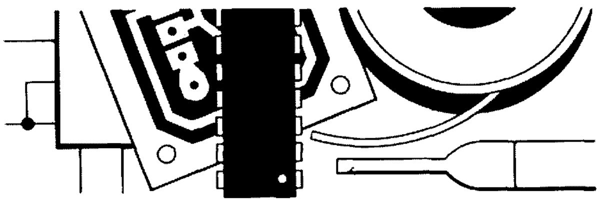
the weekender



linearity meter for rf power amplifiers

In a previous article I remarked that speech splatter is almost impossible to detect from the usual transmitter meter indications.[1] This is certainly true for meter indications of plate and grid current and power output. However, a method used in commercial equipment is available for obtaining a meter indication of splatter, which can be added to any linear amplifier as an external "black box." This article describes the design and construction of such a meter and gives hints on its use.

linearity constant

Recall that a linear amplifier, by definition, obeys the relationship $P_{out} = k\,(P_{in})$, where P_{out}, P_{in} are output

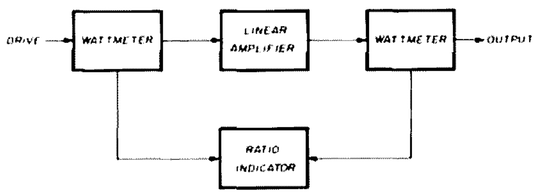


fig. 1. Principle of the linearity/splatter meter. The ratio P_{out}/P_{in} is determined. As long as it remains constant operation is linear and no splatter is generated.

and input power and k is a constant, which is determined by amplifier design and which does not change during amplifier operation. In practical linear amplifiers k is most often about 20. But it may be as low as 5 to 7 or as high as 50 to 100. The lower values are usual in grounded-grid designs; the higher values prevail in class AB_1 designs using tetrode tubes.

*A complete parts kit for this ssb linearity meter is being made available in conjunction with this article. For ordering information and prices, write to G.R. Whitehouse & Company, 10 Newbury Drive, Amherst, New Hampshire 03031.

By R.P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

Reference 1, in discussing sources of splatter, noted that k could increase or decrease, with splatter generation occurring for either case. An increase in k is associated with instability: for example, with a parasitic that is keyed on, at, or near modulation peaks. A decrease in k is almost always the result of flat-topping (driving the amplifier to the point of saturation).

meter design

These factors point the way to the design of a splatter meter. Simply put, this means a circuit that measures k. As long as k remains constant the amplifier operates linearly and will not generate splatter. However, if k changes by an appreciable amount, the amplifier is not linear and is probably generating splatter.

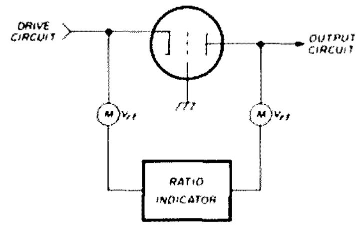
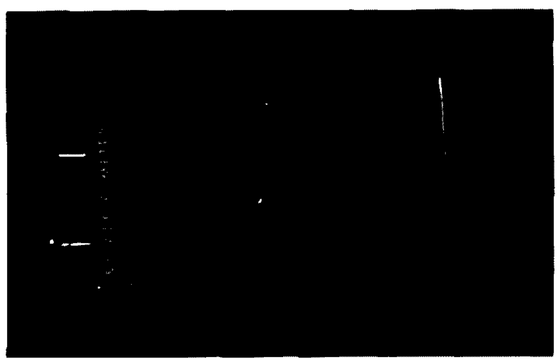


fig. 2. Linearity meter using voltage measurement applied to a grounded-grid amplifier. See ARRL handbook. Circuit must be built into amplifier.

One method of designing such a meter is evident from the equation above. As shown in fig. 1, two wattmeters are used; one measures the drive power to the amplifier and the second measures its output. The ratio of these meter readings is the desired indication. Those amateurs who have two wattmeters can connect them in in this fashion, using the simple ratio circuit described later to obtain splatter-detection capability.

In most situations it's not necessary to measure the powers. Instead only the circuit voltages or currents

Completed linearity meter for measuring the splatter of single-sideband signals.



need be measured; usually the voltages, since this is easier. Such a splatter meter is sketched in **fig. 2** for a typical grounded-grid stage. Design values for this circuit are in the ARRL *Radio Amateurs Handbook*, Chapter 6 of the 1972 edition. (The circuit is not in the 1975 edition, but may be in some earlier ones.)

Owners of the Collins 30L1 amplifier will recognize this circuit as the heart of the *load* position of the meter indicator. In this amplifier, k was determined during design and operating instructions call for the loading to be adjusted to give this value. The circuit can also serve as a nonlinearity or splatter detector, but the presence of automatic load control makes this a secondary use.

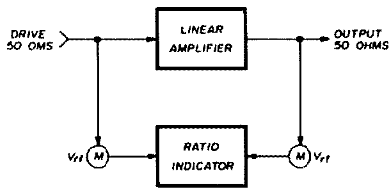


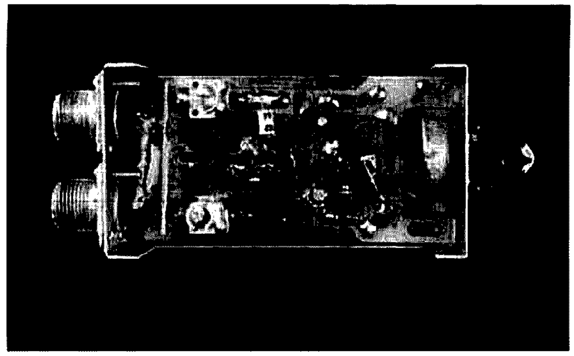
fig. 3. External linearity meter, which measures voltages at 50 ohms. The circuit can be applied to any linear amplifier.

The circuit of **fig. 2** can be added to any amplifier. Most users of commercial amplifiers don't want to make changes to the equipment lest they reduce the resale value. Also, many amateurs seem to prefer separate black-box add-ons for such functions.

practical circuit

The splatter meter is easily made into a separate add-on circuit by changing the design voltage levels of the input and output voltmeters to agree with those of the amplifier in use. For most designs, this means measuring the voltage across the 50-ohm input and the 50 ohm output, as in **fig. 3**.

Suppose that the amplifier is a typical one-kilowatt linear, with about 50 watts of drive requirement and about 500 watts output. The input voltage is about 50 volts rms (70 volts peak), and the output voltage is $\sqrt{10}$ times the input, or 158 volts rms and 221 volts peak. The voltmeters should have some adjustment range, since a) the impedances may not be exactly 50 ohms, and b) amplifier output tends to fall off on the higher-frequency bands. An adjustment range of 3:1 or so allows for this.



Prototype 50-ohm splatter meter using circuit of fig. 4. The zero-center meter is a surplus item, originally intended for fm, a-m tuner use.

One such circuit is shown in **fig. 4**. Two capacitive dividers reduce the voltage on the input and output lines to a common working level, in the range of 2 to 5 volts peak: this is the main factor in setting the ratio value. Germanium diodes give the corresponding dc levels. The meter has a zero center, with switching arranged to give E_{drive}, E_{output}, and $E_{output} - E_{drive}$. The first two are for convenience in setting up the instrument, and the last is the desired linearity measurement, or splatter detection.

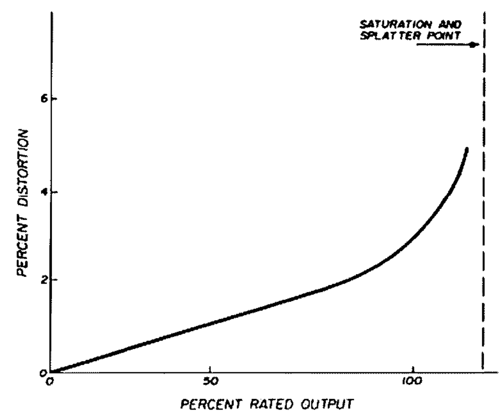


fig. 5. Normal performance of a linear amplifier. Distortion is approximately 30 dB down at the rated output. An actual curve may show more variation from a straight line.

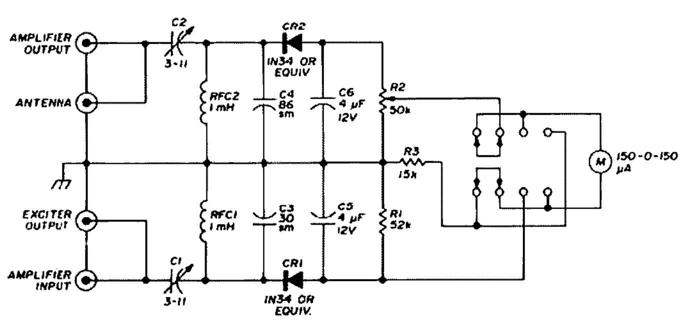


fig. 4. Circuit for an external (50 ohm) linearity/splatter meter. C1, C2 are set for power levels involved. R2 compensates for changes in antenna load and amplifier efficiency as band is changed.

construction

Except for good shielding between the drive- and amplifier-output terminals, circuit construction isn't critical. The PC board layout is shown in fig. 6. Note that the PC board is used as a part of the shielding. The drive side connectors, choke, and variable capacitor are on one side of the board and the output side elements are on the other.

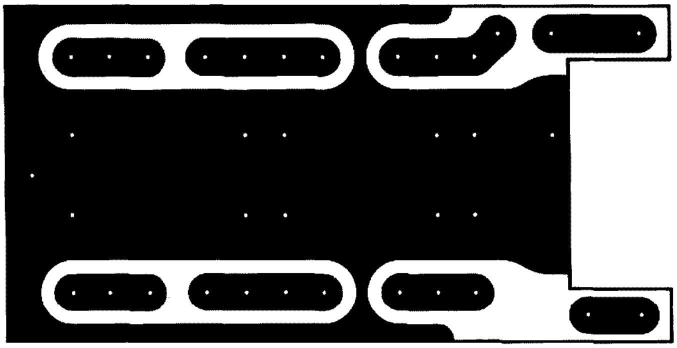


fig. 6. Full-size circuit-board layout for the splatter meter. Component placement is shown in fig. 7.

operation

Set the potentiometer to half scale and C1, C2 to minimum capacitance. Set the transmitter for half maximum output voltage (¼ power) using a dummy load. Place the meter switch on drive voltage and adjust C1 for half-scale meter reading. Adjust C2 with the meter on output voltage. Switching to the differential position gives the departure from linearity. The circuit sensitivity increases because the voltage-dropping resistor, R3, is out of the circuit in the differential position. When the band is changed, it may be necessary to reset the potentiometer for reasons discussed above.

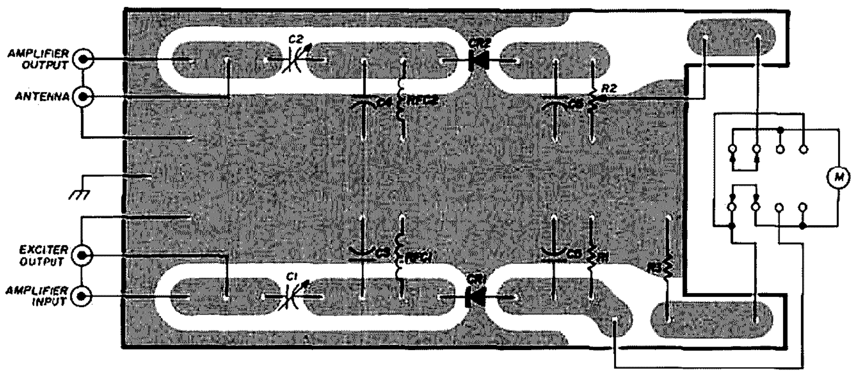


fig. 7. Component placement for the splatter meter circuit board.

Full-scale meter deflection should correspond approximately to a 10% departure from linear operation, or to distortion products 20 dB down. This sensitivity makes it easy to keep operation well below the point of objectionable splatter. Negligible loss of amplifier output occurs. The distortion-output curve of most amplifiers behaves as shown in fig. 5. The difference between a reading just discernible on the meter and bad splatter amounts to a few tenths of a dB.

A few tricks make the splatter meter easier to use. One is to use a matchbox between the antenna and the amplifier, which eliminates a variable because the amplifier output will be presented to a constant impedance. Another trick is to bias the readings slightly. I find it easiest to set up for a small, just-perceptible deflection to the right under tune-up conditions. Drive is then increased until the meter just barely flicks to the left; overdrive is indicated by a larger deflection to the left. Amplifier instability shows up as a large swing to the right.

I've made no real effort to optimize the design. The relative sensitivity on the linearity position and the time constant of the meter filters can probably be improved, as can the internal layout.

reference

1. R.P. Haviland, W4MB, "Single Sideband Speech Splatter," *ham radio*, September, 1975, page 28.

ham radio

improving transmitter keying

Methods for shaping code characters to eliminate key clicks

During a decade of activity in the ARRL intruder watch program I have observed that insufficient attention has been given by amateurs to the keying quality of CW and RTTY transmitters. This article addresses the problem of keying transients in terms of keying-circuit time constants and presents several methods for optimizing these time constants to eliminate this form of interference to other stations.

A properly designed and operated radiotelegraph transmitter produces radiation in the form of clean-cut dots and dashes having rounded edges. If the energy contained in these dots and dashes rises and falls too rapidly, the abrupt changes in amplitude cause high-order sidebands, which produce energy at a frequency greatly different from the transmitter-carrier frequency. Although weaker than signals at the carrier frequency, the sidebands introduce disturbances in the form of clicks or thumps in nearby (or even distant) receivers, even though these receivers are tuned to other frequencies. Such transients may be eliminated or greatly reduced by proper design of keying-circuit time constants.*

rise time

Observation of signals indicates that the rise and fall time of dots and dashes should be at least five milliseconds long – after passing through all the following linear stages. This rise and fall time is satisfactory at code speeds through at least 60 words per minute. **Fig. 1A** shows a dot with rise and fall times of about one millisecond, which causes key clicks in a distant receiver. **Fig. 1B** shows the dot with the rise and fall times extended to five milliseconds, which provides a clean signal that will not cause interference in a distant

*One equipment manufacturer points out that propagation conditions are the cause of some key clicks. However, this leads to a misconception because a wave shape that is abrupt may generate clicks which may not be noticed on weaker signals or under other conditions. One phenomenon that brings out key clicks probably is selective fading which may permit independent fading of frequencies as close as 500 Hz. This is noted when a carrier fades down in a-m reception, leaving the overmodulated sidebands. Similarily, a CW signal can fade, leaving the adjacent key clicks which stand out as spikes on an oscilloscope display of the received signal.

By Bill Conklin, K6KA, Box 1, La Canada, California 91011

receiver tuned to an adjacent signal while using a 200-Hz CW filter.

The dot shapes shown in **fig. 1** are those displayed on an oscilloscope connected to the audio output of a distant receiver, with agc off, and rf gain turned down appropriately.

contact bounce

Contact-bounce clicks are generated in almost all types of keys and relays. An electronic keyer may be completely free of them provided no relay is used. An example is transistor keying in the grid circuit. For some years I have used a 2N398B pnp transistor driven by a 2N404 pnp transistor at the end of a discrete-component keyer.[1] The 2N398B operates on the negative 25 volts from the grid-keying lead of a Collins 32S3 exciter. The self-completion feature of the keyer leaves nothing to create a contact bounce in the keying circuit.

The Curtis keyer kit[2] uses a 2N4124 npn transistor driver and a MJE350 pnp transistor switch for negative

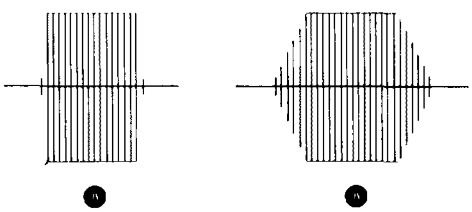


fig. 1. Dot waveforms at 1 kHz beatnote showing effect of keying-circuit time constants. Abrupt dot (A) has less than 1 ms rise and fall time, generating key clicks. Dot at (B) has a minimum 5 ms rise and fall time.*

keying voltage (or a 2N5656 npn transistor for positive keying voltage).† The Curtis 8043 keyer IC has built-in debouncers for each keying line. These are proprietary circuits with gates that provide feedback in conjunction with RC timing networks, apparently so arranged that the addition of the two paths tend to fill in any "holes" caused by bounce.

A dot with contact bounce on make looks like **fig. 2** in a distant receiver. Most bounce occurs in the first three milliseconds, although a few keyers that use keying relays have actually split a dot into two halves. Also, bounce can occur at the break end of a dot.

*Rise and fall time can be measured by displaying on an oscilloscope the output of a receiver tuned for a 1-kHz beatnote, then counting the "cycles" (which are one millisecond each). When using this simplified measurement technique disregard any low-level cycles adjacent to the baseline.

†The labels for negative and positive keying were inadvertently reversed in the keyer schematic in reference 2. However, author Olsen clearly states that an npn device must be used for positive key-up voltages and vice versa.

RTTY bounce

In manual keying most bounce is near the front end of the character, although some bounce on break is possible. In RTTY keying, bounce is more frequently

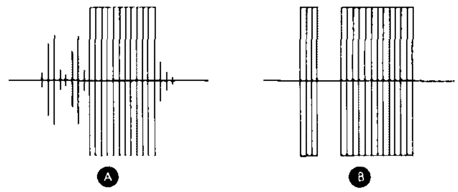


fig. 2. The irregular bounce at (A) looks like noise and is associated with "mushy" key clicks. The square-shaped bounce in (B) may not be noted by ear, except that the abrupt rise and fall times generate sharp key clicks.

encountered with keyboard operation and less so on tape with the transmitter-distributor that has a commutator. The mark pulses as seen on the oscilloscope usually start with the bounce, while the space pulses end with it. The bounce causes a fast reversal between mark and space and back again, which causes fast, broader clicks. Such conditions may create variable distortion at the receiving printer. Cleaning the contacts and varying the voltage across the contacts may change the pattern but may not entirely eliminate the problem.

vox clicks

The adverse effects of vox-type clicks and possible contact bounce can be reduced materially. It helps to have a long rise time provided by some form of RC filter in the vox circuit. If the rise time is extended to ten milliseconds, the signal amplitude at the time of the

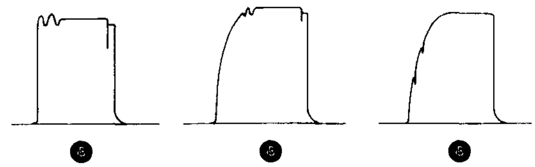


fig. 3. Key output dc waveforms showing bounce at both ends of a dot (A); with 5k resistor in keying line (B); and after adding a 0.1 μF capacitor across key (C). Bounce on make is barely perceptible (well down the curve); on break it has disappeared.

vox-type click or at the end of the bounce will be considerably less than that of the full exciter output.

key capacitor

Fig. 3A shows a dot keyed by a reed relay in a Palomar electronic keyer. Note the bounce at both ends of the dot. Any part that exhausts itself in the fully on or the fully off condition in the exciter may be expected to cause no trouble. However, it's possible to add further external treatment.

Although my Collins 32S3 vox circuit won't operate with much more than 5k ohms *added in the grid-keying* lead, it will take a 0.1 μF capacitor from key to ground. Without this capacitor, a 5k resistor between the key and exciter produces the RC time-constant curve of **fig. 3B**, but the bounce is not affected. When the 0.1 μF capacitor is added the bounce moves down on the slope of the curve and the bounce on break disappears, as shown in **fig. 3C**. Just which method is better depends upon where on the curve the transmitter output is interrupted by the bounce. After experimenting with debouncing circuits, it appears that the capacitor always should be across the key contacts.

Kenwood example

Problems with click elimination in a T-599 Kenwood transmitter may be of interest. All attempts to insert adequate resistance in the keying circuit unacceptably reduced rf output. At a distance of about one-half mile (1km) from the transmitter, I watched the keying patterns with an oscilloscope on my receiver while changes were made. The clicks were from contact bounce, and were not caused by the T-599 at all!

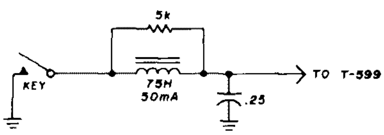


fig. 4. Inductive filter used on Kenwood T-599 to eliminate contact bounce and provide rise and fall time.

A 75 henry, 50 mA audio choke with a resistance of 2300 ohms was installed in the key lead instead of a resistor. To reduce contact bounce, a 0.25 μF capacitor was placed from the key line at the transmitter end to ground. This circuit resulted in excellent rise and fall time with no bounce, but the dots were shaped like dumbbells! A potentiometer was tried across the choke and was finally replaced with a fixed 5k resistor. The final circuit is shown in **fig. 4**. It wasn't necessary to put another capacitor across the key and no separate debouncing circuit was necessary.

The result was a keying waveform with a rise time of 6 milliseconds, a fall time of about 3 milliseconds, and a flat top similar to **fig. 1B**.

debouncing

Debouncers can be applied to keys and relays and can be used in RTTY terminal equipment.[3,4] They generally use a one-shot IC, such as the 74122 and the dual 74123. Several debouncers provide gates to use the switch or key circuit and a one-shot output (74122 or 74123 with the feedback connection to prevent retriggering).[3] The keyed output is delayed for at least five milliseconds while the bounce has a chance to settle down, with or without a later re-extension of the key-down pulse to restore its original length.

This approach appears to be good for calculator keyboards, flip-flop testers, and many other applications in which the final clean pulse length is not highly important. Note, however, that loss of a few milliseconds at the start of a dot is not noticeable to the ear, although it may change the best setting of the range

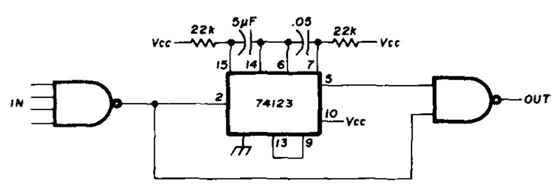


fig. 5. A type of delayed start circuit to sliminate bounce on make when contacts are closed.

selector in a teletype machine. A dual one-shot (74123) circuit is shown in **fig. 5**.

Another debouncer method is to use a gate to add the key pulse and a one-shot output during the first five milliseconds of the key-down position, thus swamping out the contact bounce on make with the one-shot output (**fig. 6**). Note that the break end is not corrected.

NE555 timer circuit

The NE555 and the dual NE556 timers simplify the circuit by eliminating the need for gates. A resistor can be brought from the 5- to 15-volt V_{cc} to the trigger input; the input to the IC simply can be keyed to ground. On the NE555 timer output end it's not necessary to add the V_{cc} side of the key to the timer output. The NE555 not only can provide a pulse of at least 5 milliseconds long (and much longer if desired), but it also can remain on as long as the trigger input (key) pulse is low (grounded).

The timer triggers on the negative-going edge of a low-going pulse, such as key down to ground. Normally

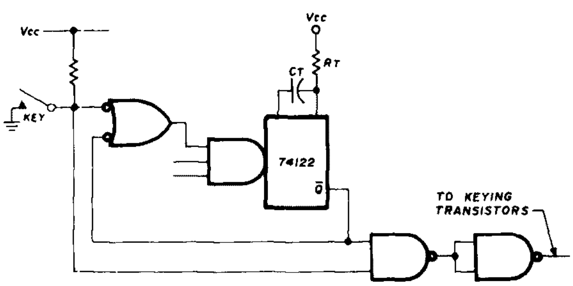


fig. 6. Circuit for swamping bounce on make by adding key output to that of a one-shot.

the trigger pulse length must be shorter than the RC time interval of the timer. If the trigger is held low, the timer output will stay high until the trigger is driven high again (key-up condition). A simple debouncing circuit for keying purposes or, for that matter, for many other purposes as well, is shown in **fig. 7**, using a common

negative isolated from ground. The V_{cc} can be 5-15 Vdc and need not be highly filtered or regulated. The output is a perfect copy of the keying but without bounce on make. It has no special provision to eliminate possible retriggering on break; but the key voltage, after rising and releasing the timer, must pass through two widely spaced voltages to retrigger the timer. Retriggering has not been observed as yet, so there has been no need to inhibit the timer for a short time at the end of a dot.

The timer output may be connected to the exciter keying input without introducing any relay contacts as shown in fig. 7 for negative grid keying, or as in the reference 2 drawing (with output polarity corrected) for positive keying. Note that internal common is *not* grounded to external ground.

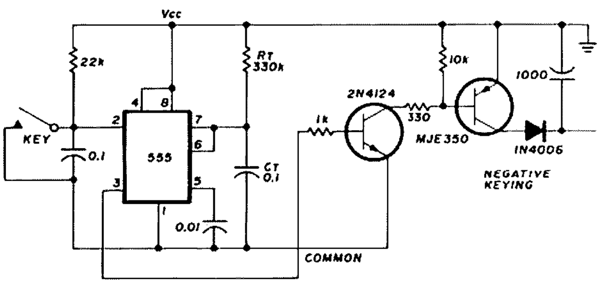


fig. 7. NE555 timer IC used to produce bounceless square output to a transistor keyer.

The keying transistor collector output can be protected against incorrect voltage polarity with a 1N4006 diode. If a ferrite bead is added for rf protection, the *large* Amidon bead, with at least 8 turns of small wire wound through it, makes a better choke than a single small bead slipped over the output lead. Whether the bypass capacitor should be on the output or on the transistor end of the bead for best rf isolation might be considered. Shielding and ground-strapping the NE555 timer and keying transistor enclosure should ensure isolation from a nearby high-power rf amplifier.

While the NE555 circuit removes the contact-bounce problem, it does have a squarewave on make and break (100 ns each). Therefore, it's still necessary to have circuitry in the exciter or, alternatively, between the keying transistors if needed, to provide a keying rise and fall time of at least 5 milliseconds for Morse keying. A suitable transition time must also be provided in RTTY keyers.

references

1. E. L. Klein, W4UHN/WB2PKE, "Der Kleiner Keyer," *73*, September, 1965, page 7.
2. Hank Olson, W6GXN, "Integrated-Circuit Electronic Keyer," *ham radio*, April, 1975, page 8.
3. John Gearhart, WAØAQO, "Scanning With a Synthesizer," *73*, April, 1975, page 23.
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ham radio

CIRCUITS & TECHNIQUES

Jim Fisk, W1DTY

coaxial connectors can generate rfi

According to a recent report from the Naval Research Labs, weak-signal communications systems can be seriously degraded by Intermodulation Generation (IMG) introduced by coaxial cable connectors which contain *small* amounts of ferromagnetic materials. Many vhf, satellite, and EME operators who use receivers with sensitivities in the −120 to −140 dBm require minimal IMG for maximum sensitivity.

NRL investigators have found that even small quantities of ferromagnetic materials in coaxial connectors can degrade IMG on the order of 50 dB. Among the connectors that can cause rf nonlinearity problems are low permeability (<2) stainless-steel connectors and connectors which have been merely plated with nickel. Hermetically sealed connectors with *Kovar** conductors are especially bad in this respect, but few amateurs use this type of connector. Since nickel plating is widely used as a substitute for silver- or gold-plated finishes, and many newer-type, low-permeability, stainless-steel connectors qualify under MIL-C-39012B, it is becoming increasingly difficult to obtain rf connectors which are completely free of ferromagnetic materials.

In the measurements made at NRL, it was found that standard, silver-plated type-N or BNC connectors and adapters are quite linear (providing the devices are clean and the contact surfaces are well aligned), allowing IMG measurements to −140 dBm (60 watts drive). Even when sixteen UG-29B/U (double female) and UG-578/U (double male) type-N connectors were connected in series, IMG performance was degraded less than 4 dB. Multiple BNC connectors were only slightly worse.

In comparison tests with hermetically-sealed connectors which use *Kovar* conductors, the IMG level typically measured −85 dBm, an interference increase of 55 dB over the −140 dBm linear device reference. To verify that the high IMG level was a function of the ferromagnetic material, a solenoid dc field was applied coaxially to the connector. The initial IMG level of −85 dBm decreased to −105 dBm — upon removal of the external magnetic field the IMG level returned to −85 dBm.

Stainless-steel coaxial connectors fared only slightly better in the IMG tests. Six double-female adapters with stainless-steel outer conductors and gold-plated, beryllium-copper center contacts measured from −85 to −90 dBm IMG. To confirm that the non-linearity was in fact due to the stainless-steel outer conductors, identical structural elements were machined from brass and assembled with the gold-plated, beryllium-copper inner contact element. With this change the IMG levels measured from −137 to −140 dBm. When the connectors were reassembled with the stainless-steel body elements, the IMG levels once again increased to −85 to −90 dBm.

To cut cost, many rf coaxial connectors now on the market are nickel plated. Unfortunately, nickel-plated rf hardware shows large non-linear rf effects. Sample nickel-plated brass devices tested by NRL have shown IMG degradation in excess of 10 dB and often higher than 30 dB relative to silver-plated brass connectors. Furthermore, the IMG performance of some heavily nickel-plated connectors was as poor as that measured with stainless-steel connectors.

In conclusion the NRL investigators recommended that immediate steps be taken to eliminate the use of nickel plating as well as other ferromagnetic materials in rf connectors, and to eliminate ferromagnetic materials from rf components where the material is subjected to the effects of current flow. They also recommended that the pertinent military specifications be revised accordingly.

Although the communications receivers used by amateurs who operate primarily on the high-frequency bands are not sensitive enough to be seriously affected by IMG produced by ferromagnetic coaxial connectors, these connectors can cause interference to other radio services, especially television, OSCAR communications, and other long-distance vhf communications systems. In addition, the NRL measurements were made at an rf power input of 60 watts — with kilowatt inputs IMG interference can be expected to be much worse.

*The thermal expansion characteristics of *Kovar* are similar to those of glass and ceramic, so it is widely used for glass-to-metal seals in vacuum tubes, semiconductors, and hermetically-sealed connectors. However, *Kovar* has high resistivity and permeability so rf losses at the seals are high, and at hf and vhf cracking can result. One amateur who used this type of connector in a 1000-watt two-meter linear lost an expensive 8877 when the ceramic seal cracked and allowed rf energy to arc to ground. As further proof of the unsuitability of these connectors for rf work, tests aboard the *USS Bunker Hill* indicated that third harmonic distortion from high-frequency Navy transmitters increased about 20 dB when hermetically sealed adapters were used.

This NRL report has widespread implications for the entire electronics industry. Television sets in fringe areas which are located near amateur or broadcast transmitters, and use ferromagnetic coaxial connectors, are likely to be subjected to undesired interference. Vhf-fm systems with close-spaced, diplexed antennas using ferromagnetic connectors may have increased intermodulation problems, and high-quality, commercial-grade antennas which feature stainless-steel hardware may have undesirable IMG characteristics.

improving speech intelligibility

During the Apollo moon shots the NASA flight controllers sometimes had difficulty separating an astronaut's voice from the background noise. Since the voice spectrum between 40 and 900 Hz and 1900 and 3000 Hz seems to contribute little to intelligibility, and only three portions of the speech spectrum are apparently

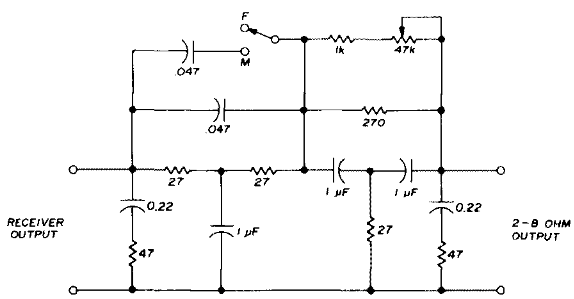


fig. 1. Circuit developed by NASA engineers to improve the intelligibility of voice communications. Potentiometer adjusts null to 600 Hz, and switch provides for either male or female voices.

required for clear speech (300-400 Hz, 900-1700 Hz [1100-1900 Hz for females], and 2500-3000 Hz), NASA engineers developed the circuit of fig. 1 for more reliable voice communications. In this circuit the potentiometer adjusts the null to about 600 Hz, and the switch provides for either male or female voices. Although the audio bandwidth of most amateur transmitters is limited to 300-2400 Hz, this circuit has been used to good advantage for pulling weak DX signals out of the noise and correctly identifying their callsigns.

low-level, high impedance preamplifier

A low-cost, transistorized preamplifier for oscilloscopes and meters can be a real time saver in the lab. The unit in fig. 2 uses a Darlington circuit to obtain extremely high input impedance. The ac input impedance of this circuit is approximately $(h_{fe})^2$ times the emitter resistance of transistor Q2, and in practice has been measured in excess of 2.2 megohms. With the input shorted, the noise level is -78 dB down as read at the output with a vtvm. By using low value resistors in the base of Q1 to establish the operating point, the circuit exhibits good dc stability over wide temperature excursions. In addition, linearity is within 1.5 percent from 100 microvolts to 1 millivolt input, and frequency re-

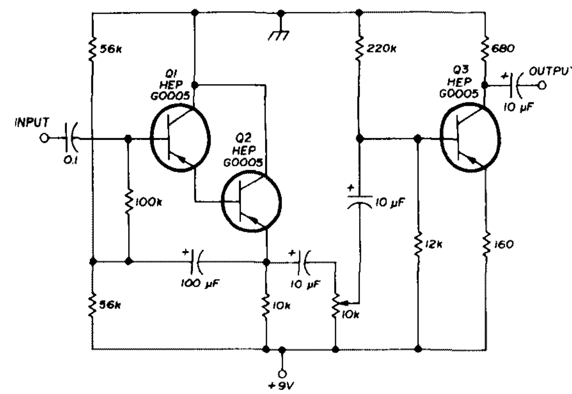


fig. 2. Low level, high impedance preamplifier has input impedance greater than 2.2 megohms. Frequency response is from approximately 100 Hz to 350 kHz. All transistors are Motorola HEP G0005.

sponse is ±2 dB from 100 Hz to 350 kHz. Although originally designed to boost the input to an oscilloscope, this circuit may be adapted to other applications where high gain, low noise and high input impedance are required.

dynamic microphones

Not too well known among amateurs is the fact that dynamic microphones exhibit a *proximity* effect which increases the bass output as the distance between the source of sound and the microphone is decreased. Although stage performers often use this effect to increase the bass registers or to add "warmth" to their voice, amateur operators should refrain from getting closer than about 4 inches (10cm) from a dynamic microphone.

The result of a signal source 3 inches (7.5cm) from a typical dynamic microphone is graphed in fig. 3. At 200 Hz the output is approximately 4 dB higher than at 1 kHz, an increase of about 58 per cent. What fig. 3 doesn't show is that the power output per unit bandwidth is considerably higher at frequencies below 500 Hz than it is at the higher audio frequencies which are more effective for radio communications. In addition, the increased bass output caused by speaking too close to a dynamic microphone can introduce distortion be-

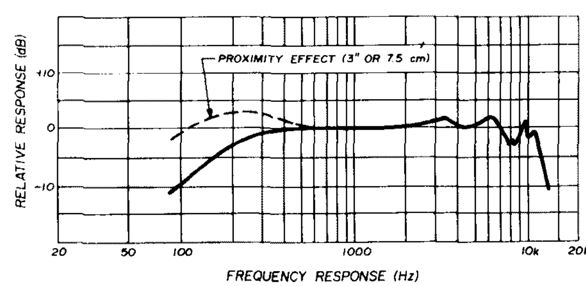


fig. 3. Proximity effect of dynamic microphones provides increased low-frequency output when the sound source is too close to the microphone. For best results in radio communications, keep the microphone at least 4 inches (10cm) away from your mouth.

cause of overloading and, in fm systems, overdeviation.

Speaking across the microphone may help, but different brands of microphones have different off-axis response and they are not very directional at the lower audio frequencies. For best results, speak directly into the front of the microphone, but keep it at least 4 inches (10cm) away from your mouth.

lightning protection

Most amateurs make sure their antennas and towers are well grounded for lightning protection, but sometimes forget that lightning can enter the service entrance to their homes, causing a good deal of damage. Since the high-voltage surges enter the service entrance and seek the least resistance path to ground, all to often that path is through your carefully grounded amateur equipment. Television sets, electric stoves, hot water heaters, and house wiring also fall prey to service entrance lightning and many bad fires have been caused by it. In most cases the damage isn't caused by a nearby lightning strike, but one to the power line a good distance from the house.

Fortunately, there is a very inexpensive solution to this problem in the form of a small, low-cost lightning arrestor which is similar to the large units used at electrical substations. Both General Electric and Westinghouse manufacture these devices. The General Electric 9L15CCB007 home lightning protector, for example, costs about $12.00, can be installed at the service entrance or inside the main breaker panel, and is guaranteed for minimum life of ten years. Some insurance companies will give a rate reduction with a properly installed service entrance lightning arrestor. With an arrestor of this type and a properly grounded antenna, the only worry is a direct hit on the house (which can be protected with lightning rods).

oscar antenna

The antenna shown in fig. 4 was suggested by WA4DDH for use with Oscar 7, both Mode A and Mode B. The antenna consists of delta-loop beams for ten and two meters, a multi-element Yagi for 432 MHz, and is

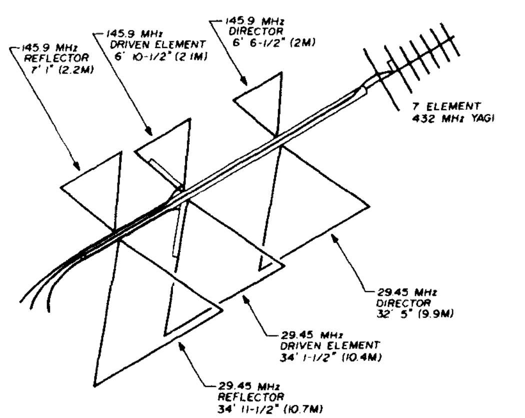


fig. 4. Antenna for Oscar 7, mode A and B, suggested by WA4DDH uses delta loops for 29.45 and 145.9 MHz, and a 7-element Yagi for 432. Each antenna has its own gamma match and transmission line.

tilted about 35 degrees above the horizon. Separate gamma matches are used on the two delta loops, with all three feedlines taped along the boom. Although WA4DDH suggested the use of a 7-element Yagi, a number of suitable 432-MHz antennas have been described in the amateur magazines.

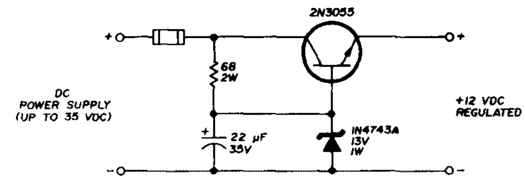


fig. 5. Simple 12-volt regulated power supply for up to 2 amperes. The 2N3055 is mounted on small U-shaped heatsink made from 2 inch (5cm) strip of aluminum.

coaxial cable

A number of readers have written to complain that coaxial cable they've purchased recently doesn't have the braid coverage it used to have. In some cases the braid coverage is so poor that you can see the inner insulation through gaping holes in the braid. This sad state of affairs is due to the fact that RG-8/U is no longer a Mil Spec cable (having been replaced by RG-213/U). Copper is expensive, so the manufacturers have cheapened the construction of non Mil Spec coaxial cable rather than cutting the price. Therefore, it's a good idea to stay away from commercial quality coax, insisting instead on those cables which are covered by military specifications (RG-58C/U, RG-59B/U and RG-213/U for example).

An additional benefit of Mil Spec coax is that the outer jacket is of the noncontaminating type. The outer jacket comes in two types and is very important in determining the life span of the cable. The older class 1 contaminating jacket (used on RG-8/U, RG-58/U and RG-59/U) incorporates a plasticizer during the manufacturing process to keep it flexible. As soon as the cable is jacketed, however, the plasticizer starts to leach through the shield braid into the polyethylene insulation around the center conductor. This changes the electrical characteristics of the insulation and results in increased cable losses. This process is relatively quick, and increased losses can be readily measured after only one or two years.

Later coaxial cables use a non-contaminating class 2A jacket which is long-lived, abrasion resistant, and not damaged by sunlight. Furthermore, this type of jacket is no more expensive than the contaminating type and can be directly buried for underground runs. Since it doesn't contaminate the center insulator, coax with a class 2A jacket has a useful life of ten to twenty years.

Although foam-filled cables are becoming more and more popular, the shield braid does not conform to Mil Specs, most foamed cables have contaminating type jackets, and unless the foam is gas filled to keep moisture from migrating into the foam, this type of coax is not very desirable.

For best results select only late types of coaxial cable which use class 2A jackets and leave the foamed coaxial cables to the Citizen Banders and CATV companies. The jacket type can be checked in the manufacturer's catalog (which your dealer should have). Coaxial cable with a class 2A jacket is usually no more expensive than the older non Mil Spec cables.

oscillator circuit with an emitter-follower output. The oscillator circuit consists of Q1 and Q2. Transistor Q3 is the emitter follower; Q4, also an emitter follower, insures good isolation between the oscillator and load, and provides a low-impedance output (200 ohms). Emitter-follower Q3 also provides isolation between the oscillator and the feedback amplifier.

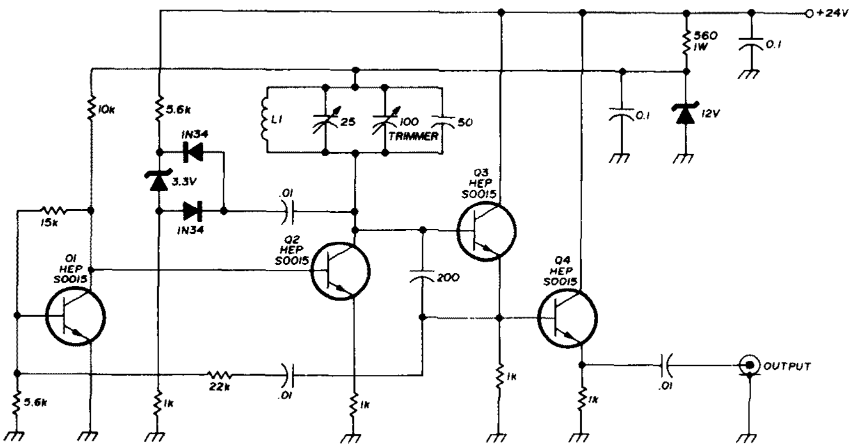


fig. 6. High-stability vfo tunes from 3.5 to 3.8 MHz. Output is 5 volts p-p minimum into a 200-ohm load. L1 is 25 turns no. 18 (1mm) closewound on a 1½-inch (38mm) diameter form. All transistors are HEP S0015 or similar.

regulated dc supply

Since many of the inexpensive 12-Vdc power supplies which amateurs use to power low-power vhf fm gear are unregulated, the no-load voltage may be 18 volts or more. During transmit the voltage drops to about 12 volts, but when receiving the supply voltage may be high enough to exceed the maximum voltage ratings of the small-signal transistors in the set. The simple voltage regulation circuit in fig. 5 will easily handle up to 2 amps or so, but a short to ground (negative lead) on the transistor or heatsink will probably burn out the transistor. One solution to this problem is to use a small plastic utility box to house the circuit. Be sure to include the 2 amp fuse – many published circuits leave out this one small but important detail. The heatsink for low-power applications (2 amps or less) need be nothing more than a 2 inch (5cm) strip of aluminum bent into a U-shape.

stable vfo

The extremely stable vfo circuit shown in fig. 6 has a total drift of less than 10 Hz from turn on, and a total frequency drift of less than 30 Hz as the supply voltage is varied from 15 to 30 volts. The frequency range is 3.5 to 3.8 MHz and output is a minimum of 5 volts peak to peak; amplitude stability over the entire tuning range is within 1 dB. Current drain with a 24-volt power supply is 50 mA.

This circuit, designed by G3ISP, is basically a Butler

The output amplitude is stabilized by the two 1N34 diodes and 3.3-volt zener. Oscillator output is coupled to the diodes through a 0.01-μF capacitor. The voltage across the zener is maintained by a resistance voltage divider, and the rectified oscillator output is compared to the fixed zener voltage.

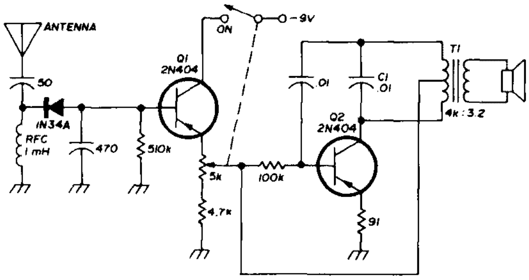
The original model of this vfo was built on perforated circuit board, 4 inches long by 3 inches wide (10x7.6cm). This was mounted in a 8x6x6-inch (20x15x15mm) enclosure along with the coil, 25 pF air variable and 100 pF trimmer. A communications receiver with a 100 kHz crystal calibrator was used for calibration.

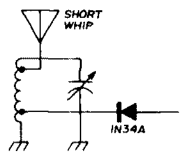
cw sidetone

Every CW operator should have an audio sidetone oscillator to monitor his fist. Whether you use a sideswiper, an automatic bug or an electronic key, a monitor is necessary to insure proper sending. The sidetone oscillator circuit in fig. 7 is simple and reliable and changes in transmitter operating frequency do not require any adjustments in the sidetone circuit.

The circuit is essentially a Hartley audio oscillator which is turned on by the diode rectifier and dc amplifier. For operation from 160 through 10 meters, the rf choke and coupling capacitor are more than adequate for coupling to the final tank coil; couple the lead just close enough to the final tank to get a four- or five-volt swing at the emitter of Q1 when the transmitter is keyed. For vhf use on 6 and 2 meters, a small tuned

circuit and pickup antenna are usually required to obtain sufficient rf to turn the monitor on. To adjust the audio tone of the oscillator output, different values of capacitance may be substituted at C1.





INPUT CIRCUIT FOR VHF USE

fig. 7. Rf-activated CW sidetone circuit can be used on 160 through 2 meters (the tuned input circuit is recommended for vhf). Audio pitch is adjusted by changing the values of C1.

transient eliminator

When the circuit of fig. 8 is placed between a dc power supply and a load, it will eliminate many transient overloads that might damage semiconductors and other components. The zener diode is chosen so that its value is slightly higher than the supply voltage so transistor Q1 is normally turned off. Transistor O2 is normally conducting and transmitting power from its emitter to collector. When a voltage spike or transient is present on the input line, the zener diode conducts and turns Q1 on. With Q1 conducting, the flow of current into the emitter places a positive bias across the 10k pot. This bias turns off Q2 during the transient and protects the load circuitry connected across the output of the protection circuit.

staircase generator

A square wave may be converted into a staircase voltage output by the simple two-stage feedback transistor network in **fig. 9**. The amplitude of the staircase voltage is linearly proportional to the number of applied pulses, and each step approximates the amplitude of the input pulse. The circuit functions in the following manner: When the first pulse is applied to the input, capacitor C2 charges through C1 and Q2 to a voltage approximating the amplitude of the input pulse. When the input pulse terminates, the voltage across C2 acts as a signal for Q1, forcing C1 to charge to the same voltage that is across C2. When the second pulse is applied, it is added to the voltage across C2, thereby causing C2 to double its charge. Each subsequent pulse increases the height of the staircase until it reaches the level of the supply voltage.

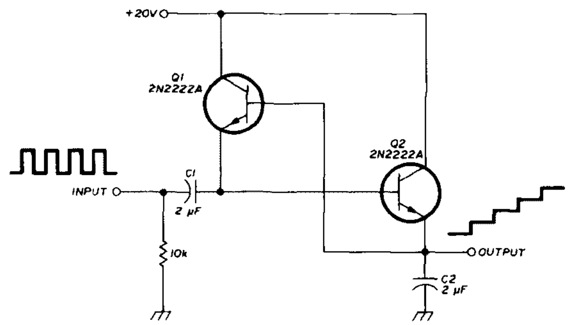


fig. 9. Simple staircase generator circuit converts a square wave into a staircase voltage output. Staircase is linearly proportional to the number of applied pulses, and each step approximates the level of the input pulse.

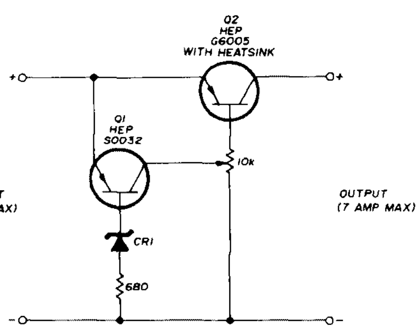


fig. 8. Protection circuit eliminates transients and voltage spikes from the supply line. Voltage rating of CR1 is chosen about 10% higher than supply voltage. The 10k pot sets input voltage at which Q2 is cut off. Although maximum current limit of a properly heatsinked G6005 is 7 amps, a similar scheme could be used with higher power transistors.

closing comments

In a recent *ham notebook* item on muting microphones* author W6IL noted that microphone disturbance (such as late at night) can be eliminated by taping a heavy-walled cardboard tube to your microphone and speaking through it. The tube compresses the sound, resulting in increased talk power, although it may sound as though you're in a barrel. To avoid the "bottom-of-the-barrel" effect, WB2CHO suggests placing a small amount of loose cotton in the tube. This damps out the echoes, and reduces the reverberation effect (which, judging from some 75-meter signals, some operators like).

WB2CHO also has some suggestions regarding speech processing and RFI as discussed in the *comments* section of the November, 1975, issue: infinite speech processing has the same effect on tube dissipation as running your rig in key down, full output most of the time. Most amateur ssb transceivers and linear amplifiers are designed for a 50 per cent duty cycle, which normal voice does not exceed. With heavy speech clipping or quasi-logarithmic processing, this duty cycle can be exceeded, however, reducing tube life or blowing out a power supply. To be safe, reduce drive when using speech clipping.

*R. Cabanillas, Jr., W6IL, "Muting Microphones," *ham radio* November, 1975, page 71.

ham radio

improved frequency readout for the Collins S-line

Two paper dials added to your S-line or KWM-2 allow frequency to be read within ten Hz

A casual glance at the accompanying photos shows what looks like an ordinary KWM-2 transceiver. Closer inspection, however, reveals that something has been done to the main tuning dial. The graduated scales shown mounted to the main tuning control allow frequency to be read to 10 Hz on a unit designed to read out in kHz. All this was done with just two pieces of paper.

You might say, "I've a good eye and can estimate frequency to 100 Hz on my Collins." Perhaps. But you'd be estimating to one-tenth of a 0.1-inch (2.5mm) space. The reliability of such an estimate is certainly questionable.

preliminary checks

Before adding the frequency dials to your Collins equipment, the first step is to determine the amount of frequency readout error that exists. After you've accurately set the vfo dial to eliminate end spread in accordance with Collins instructions, check the 100-kHz point on the dial. You'll probably notice an error of between 100 and 400 Hz. A recheck then should be made.

Put a reference mark on the desk. To avoid parallax, align the reference mark, using a magnifying glass, with the 100-kHz point. Offset the hairline 1 kHz so you're using the thinner and more precise dial marks. You'll probably find the frequency is still off at the 100-kHz check point.

If you have an accurate marker generator that goes to 5 or 10 kHz, you can check across the entire 200-kHz range. If you plot the data you'll most likely find an error curve that's somewhat sinusoidal. However, you'll seldom see more than 50-Hz error between any adjacent 5-kHz marks, even though the total error across the 200-kHz range may be as much as 400 to 600 Hz. If you had accurate 5-kHz markers and some way to read the dial to 10 Hz or so, you could plot an error curve, and by using it, be fairly sure of the frequency of any signal to within 25 Hz or better.

applying some leverage

What's needed is a bigger dial with marks, say, every 100 Hz instead of only every kHz. If you could set the dial accurately, you could also use a vernier to read to 10 Hz (and even estimate to 5 Hz).

The Collins knob turns nearly 9 revolutions to cover the 200-kHz range. The knob has a skirt circumference of about 8 inches (203mm), so if the knob skirt is appropriately marked, you would have nearly a 6-foot (1.8m) dial. If a second dial, also appropriately marked, were added to provide a vernier measurement then you would have the equivalent of a 60-foot (18.3m) dial.

And that's exactly what was done. The dials, shown in fig. 1, were made as described below and installed on my KWM-2. The dials will also work, with slight error, on the Heath SB-301, 303, and 401. The large dial is graduated in kilohertz from −10 through zero to +10. It is mounted to the Collins transceiver main tuning shaft

By Bob Gilson, W1GFC, Muhlthaler Str. 5, 8131 Starnberg-Hanfeld, West Germany

so that it doesn't move (press fit). The center hole of this dial must be cut as accurately as possible so that no play exists between it and the tuning shaft.* The small dial is graduated with 10 marks for each 0.9 kilohertz and is, therefore, a 10:1 vernier. The small dial is marked from 1 to 26 for reference. This dial is cemented to the back of the Collins tuning knob.

resetability checks

Before you decide to install the dials, here are a few words of advice — learned the hard way: make a check for resetability and backlash in your unit.

Mark a reference line on the knob skirt and on the front of the transceiver (white adhesive tape works well). Carefully align the hairline at 200 on the dial and count the number of revolutions of the knob to near zero. You'll find nearly 9 revolutions will be required. Now tune from zero to 200 and see if you get the same reading. In my case I read 8 turns from 200 to 12 and 8 turns from zero to 188.5. That's an accumulative error of 0.5 kHz across the tuning range, and if linear, would be 12 Hz between any two 5-kHz marks, or 6 Hz from the nearest mark.

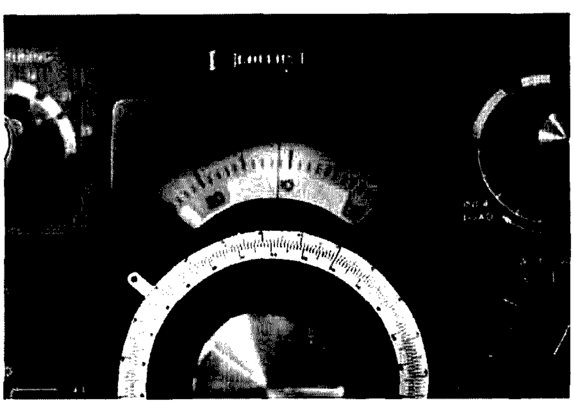
If you find that one full knob revolution covers about 23.5 kHz, the dials are for you. If you find the knob covers less than 22 or more than 25 kHz, the dials are not recommended.

making the dials

I used heavy printer's stock and mounted it on a Bridgeport milling machine table. Then I learned that graduating 3-inch (76mm) diameter dials with marks every 1 degree, 22.5 minutes was a chore. Finally I used a 15-inch (381mm) diameter dial. The circular table of the milling machine is readable to one degree of arc, and the crank reads to one minute and can be interpolated to 30 seconds. On a 15-inch (381mm) diameter dial, the

*The dials may be cut using a pair of dividers. Easy does it: rotate the dividers, using moderate pressure, until the dials start to separate from the paper. Then gently separate by bending back and forth. Clean the edges with fine-grit sandpaper. Editor

Closeup of Collins KWM-2 with frequency dials mounted. The equipment indicates a frequency of 14,211,220 Hz.



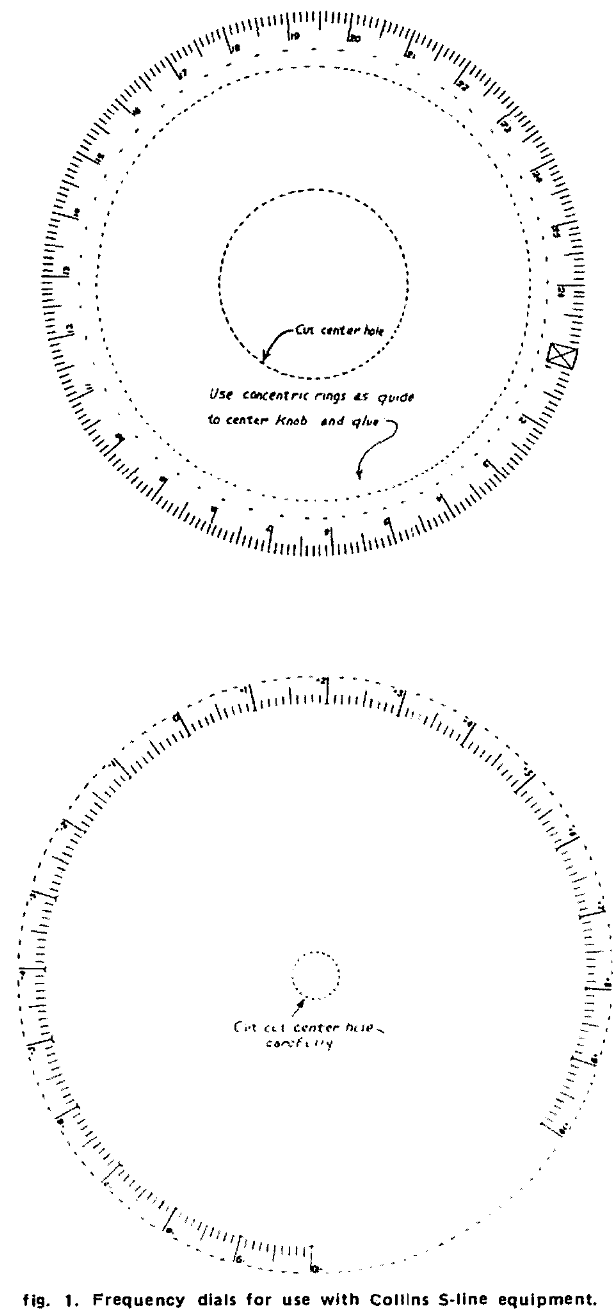


fig. 1. Frequency dials for use with Collins S-line equipment. These dials have been photo reduced from the original upsize masters, which were made on a milling machine table.

lines are 0.01 inch (0.25mm) thick. So even though the Bridgeport machine provides about 0.6 Hz resolution (in terms of frequency), the lines are 7 to 8 Hz thick. This inaccuracy, combined with occasional drafting errors, means that the average accuracy is 10 Hz or so.

The large dials were photo reduced and printed on stiff paper to 3 inches (76mm) diameter for the KWM-2. The printer did a good job — I measured 3.0010 inches (76.225mm) diameter on dial A and 3.0012 inches

(76.230mm) on dial B, using the shadowgraph method. The main thing, however, is uniformity rather than exact dimensions of the dials.

installation and use

The dials should be cut out as indicated in **fig. 1**. Use care with the small center hole on the large dial because it should make a snug fit on the Collins tuning shaft. Also use care in cutting the circumference of the small dial, as it will become the vermier. After cutting out the dials, but before mounting, check to determine correct alignment of the two scales. If, for example, you align 15 on dial B with 0 on dial A, then 10 and 20 on dial B should line up with -4.5 and +4.5, respectively, on dial A; otherwise the dials haven't been cut and centered correctly.

Remove the tuning knob (two set screws) from the shaft and slide the large dial onto the shaft. Unless you've had experience, do not remove the nut on the shaft, even though the dial isn't flush. Center the knob skirt over the small dial, using the concentric circles as a guide, then cement the dial to the back of the knob with the numbers facing out. Re-install the knob onto the shaft.

Use any accurate marker generator (10 kHz or 5 kHz if you have it) and tune to the nearest marker in the part of the band where you want to measure frequency. Tune for zero beat (or offset if you have an audio oscillator), and set the large dial to line up with one of the numbered marks (1 to 26) on the small dial. You can use 0 kHz, for example, on the large dial. In this case, the large dial would be used much like the hairline on the regular Collins dial.

Hold the large dial and rotate the knob to the signal of interest (either zero-beat or audio offset), and read kilohertz and tenths of kilohertz on the large dial. You can read hundredths of kHz on the smaller dial where the lines line up. The correct vernier lineup is the first one beyond (in the direction you moved from reference) the tenth of a kHz.

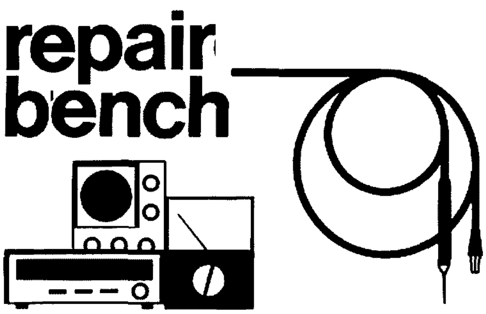
The KWM-2 in the closeup photo is indicating 14,211,220 Hz. The first number (2) of the last three digits is read from the position of reference 12 on the small dial to the right of zero on the large dial. The last two numbers (20) are shown where the vernier lines up with the large dial (two lines to the right of the 12 on the small dial).

After you've used the dials for awhile, it's suggested that you cement a tab to the back of the large dial to use as a holding tab when you turn the knob (and the other dial). If you're right-handed you may find that mounting the tab near -3 kHz will be most comfortable.

final remarks

With the dials installed I can read frequency to 10 Hz and estimate to 5 Hz. With an accurate 5-kHz marker, I feel confident that I know the frequency to 20 Hz. My objective was to read frequency accurately without a counter. I succeeded, but I had to use a counter to verify my success.

ham radio



Joe Carr, K4IPV

basic troubleshooting: the dead receiver

Many amateurs seem to view troubleshooting as some kind of arcane, unfathomable process performable only by professional electronics personnel. It is my contention, shared by many others, that most troubleshooting of amateur radio equipment is a relatively simple skill which is easily acquired by anyone with a technical background good enough to pass the General or Technician class FCC examinations. To be sure, a person with twenty years experience in a communications service shop might run rings around the neophyte timewise, but your rig will work just as well after you correct the defect as when he does that neat trick for you.

The first step in any troubleshooting procedure is to observe performance of the defective unit. The test equipment requirements for this phase are almost zero. It is at this point, by the way, that the real pro exemplifies himself. There is a great deal of information to be gleaned from this process and it can significantly reduce the time required to diagnose and repair an equipment fault. Since the pro is in the business of making money he will use this information to full extent. You, too, should make use of your observations.

In order to simplify the discussion of "troubleshooting" I have limited the range of interest to the single-conversion, superheterodyne communications receiver and, by implication, the receiver portion of the typical ssb transceiver. Of course, it takes only the slightest effort to extend the procedures given here to dual- and triple-conversion models.

When you initially examine the receiver, determine just what is or is not working. The fact that the receiver is dead, while significant and interesting, is not the whole story. For example, does "dead" mean that no power seems to be applied, that neither the pilot lamps nor the tube filaments are glowing or is it simply a matter of no output even though the lamps and filaments seem normal?

In cases where the lamps come on and the tube filaments glow, you can look for some other, more subtle clues, which can be immensely valuable. Look first to the audio gain control. If you hear a scratching sound as you rotate the control through its range then it is a moderately safe bet that the defect is prior to the volume control. What you are hearing is the amplified noise voltages generated by the control wiper.

Next, switch on the bfo or turn to the *ssb/CW* position of the function switch. If the noise level increases or there is static as you turn the switch then one can generally assume the trouble to be prior to the detector. The next step might be to switch the bandswitch through its ranges. Look for either signals or static crashes as the switch changes bands. If the receiver is not dead on all bands look to components which are only common to the dead band or bands. Such components might be coils, converter crystals in double-conversion sets, loose connections or even the switch contacts themselves. This last is often overlooked by the inexperienced. On the other hand, if static crashes are heard as you switch bands then you can temporarily exonerate the i-f amplifier chain.

In cases where the set is completely dead, as if the power switch was turned off, look to the primary side of the power supply. Actually, defects such as this, although they initially look bad, are actually pretty easy to solve. It seems that there is nothing quite so easy to find as the cause of a completely dead set. First and foremost, check to see if the ac receptacle has power and that the set is plugged into the receptacle. You would probably be surprised at how many service calls in the TV industry are due to something like a wall switch in the *off* position! Use an ohmmeter or continuity tester to check the fuse, power cord, power switch, and trans-

Technician using a modern hand-held digital voltmeter to troubleshoot electronic equipment. (Photo courtesy Hewlett-Packard)

former primary. Don't rely on sight to test the fuse – actually measure its resistance (should be zero ohms). Be aware, though, that a blown fuse may indicate some other defect which can be quite serious. An old maxim regarding fuses goes something like this, "A fuse doesn't cause trouble – it protects against and *indicates* trouble!"

Include in your preliminary inspection any unusual odors (might indicate that something is burning), any unusual sounds from the loudspeaker, any sounds from inside the cabinet which might tend to indicate arcing, and so forth. Although it has taken quite a bit of time to describe this inspection the actual implementation takes only a few moments. If you use these observations, coupled with a logical step-by-step procedure, there are few defects which will elude you in a typical superheterodyne receiver.

divide and conquer

One of the best methods for isolating defects in most types of electronic equipment is to divide the set into bite-size chunks then conquer each in its turn. There are two basic philosophies to troubleshooting, signal injection and signal tracing, but they both lead to the same thing. Once you have gained experience you will find that one will work better than the other in some cases, while in others the reverse is true. Most times, though, which to use is really only a matter of preferance so take your pick.

The signal injection procedure is shown in **fig. 1**. First connect some sort of output monitor. This can be an ac voltmeter, an oscilloscope or just a loudspeaker. Then inject a signal from a generator into the input of each stage in succession beginning with the output. Of course, if your preliminary investigation has shown the audio stages to be working then you can just skip them and go on to the detector. Check first one stage, then the stage preceding, until the stage is located which either passes no signal or a highly attenuated signal. You can then be reasonably sure that the defect is located between that point and the last point where signals passed normally.

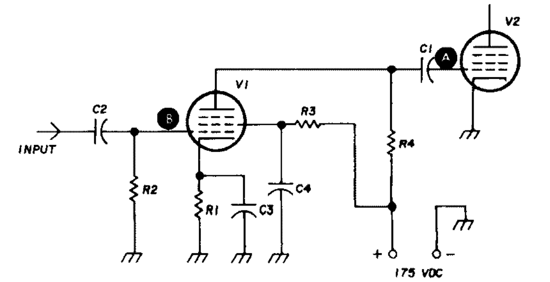


fig. 2. Typical audio amplifier stage which might be found in a vacuum-tube amateur receiver. Various faults which can be caused by component failure are discussed in the text.

Consider as an example the audio amplifier shown in fig. 2. Assume that output was heard when the signal was injected into the grid of V2 (at point A) but no output was produced when the signal was applied to point B, the grid of V1. The next step, of course, would be to check the vacuum tube either by substitution or on a tube checker. If you or a friend owns a tube tester then by all means use it but don't feel demeaned by going down to the local drug store tube tester (they work).

Since some defects will not show up on a tube tester I feel that substitution is the best method. The number of different tubes which the typical amateur is likely to have in his primary station equipment is low enough to warrant keeping one each (new, not hamfest specials of unknown history) in stock at all times. In most cases replacing the tube will cure the defect; although probably not the 95% claimed by tube tester ads.

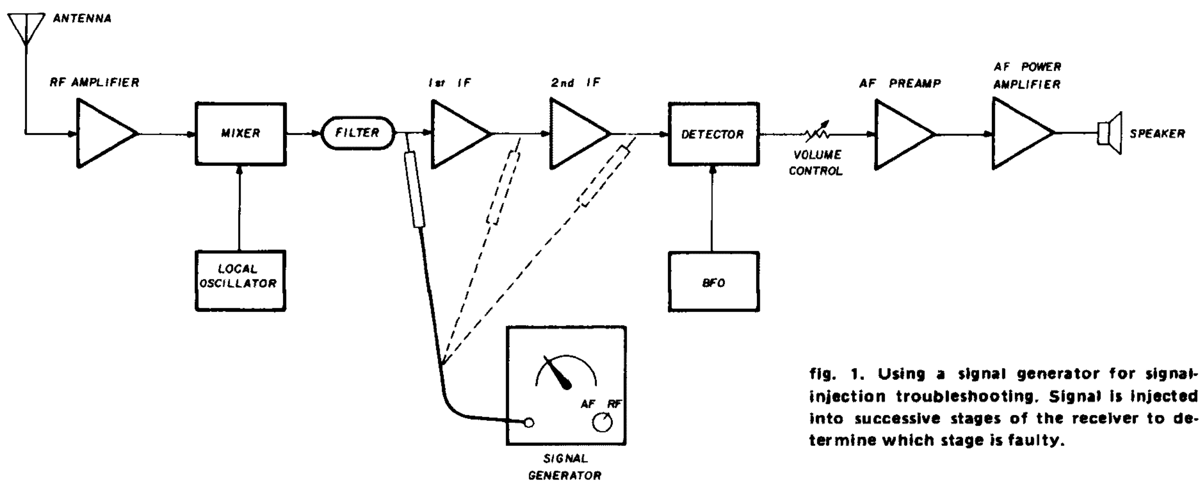


fig. 1. Using a signal generator for signal-injection troubleshooting. Signal is injected into successive stages of the receiver to determine which stage is faulty.

You all know Murphy's law (one corollary of which states that Murphy was an optimist), so in your case, alone, the trouble will not be a tube and you will have to look further. When you remove the tube from its socket, leave the set turned on unless this exposes you to dangerous voltages while attempting to extract that tube

from its socket. If you hear a "click" or static crash as the tube clears the socket pins then you know it was drawing current and can probably forget about the plate load resistor, or in rf/i-f circuits, plate tank coil.

When you cannot cure the problem by replacing a tube it will be necessary to make some voltage measurements on each pin of the tube socket. Unless you have all the tube pinouts memorized it's advisable to obtain a copy of the receiver schematic. If one is not available try getting one from a different version of the same model. Failing that, you can always consult the RCA *Receiving Tube Manual,* GE's *Essential Characteristics,* or the rear of any edition of the ARRL *Radio Amateur's Handbook*. Compare the voltages actually measured in the receiver with those given in the schematic. A variation of 10 to 20 per cent is normal.

If there is no schematic available then try comparing the voltages against what common sense tells you is approximately correct. I kept records in a shop and from them noted that nearly half of the jobs which were not tube replacements involved either the dc power to a stage or produced changes in the normal dc potentials on the tube elements. A shorted screen-bypass capacitor, for example, will drop the screen B+ to almost zero and may well overheat the screen-dropping resistor. An open cathode-bias resistor, on the other hand, will cause the plate and screen voltages to rise and the cathode-to-ground voltage will be almost equal to the plate voltage.

Note that there are actually two paths for current to flow in almost every receiver or audio amplifier circuit: the dc path just discussed and an ac signal path. When dc conditions look normal then look to that ac path for the defect. In the audio amplifier of fig. 2 the ac path includes capacitors C1, C2, and C3. Should they open the set will be dead but the dc potentials will be close to normal. Capacitor C4, also in the ac path, will generally cause the stage to oscillate if it opens up. Simple substitution using either alligator chip leads or "solder tacks" is the best troubleshooting method.

An alternate troubleshooting technique, illustrated in fig. 3 is called signal tracing. It is actually just signal injection viewed from the opposite end of the cascade chain of stages! Here, you inject a signal into the input, antenna terminals if a receiver, and look for it in successive stages using a simple signal tracer or oscilloscope. The most basic form of tracer is a high gain audio amplifier preceded by either a demodulation or low-capacitance probe depending upon which stage is being tested. Both of these probes (see figs. 4 and 5) are also needed when using an oscilloscope. Almost any audio amplifier can be pressed into service: junk public-address amplifiers, hi-fi amplifiers, a homebrew amplifier, or the audio stages of an old radio. Alternatively, you can use any of a number of small kits offered by most electronic supply houses. The only real requirement is for moderate to high gain and enough audio power to drive a small speaker.

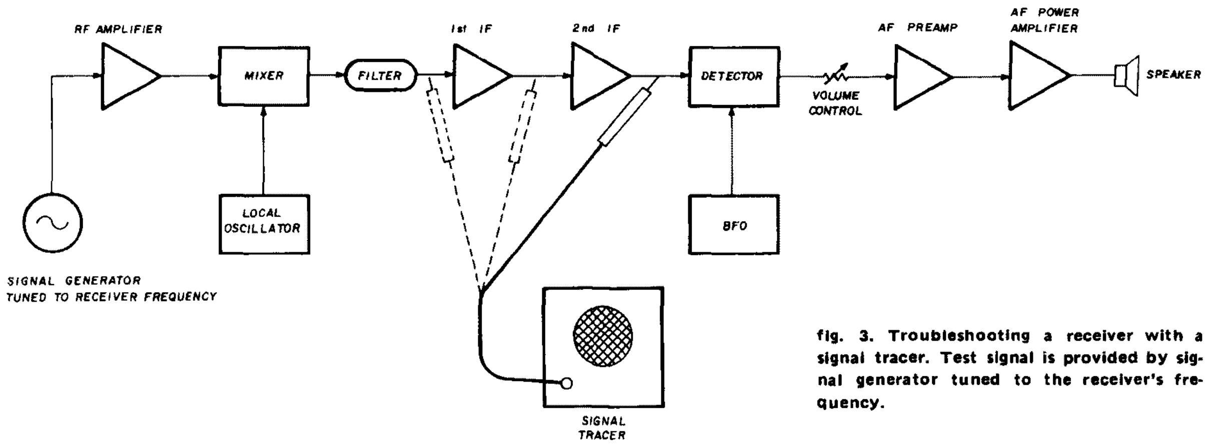


fig. 3. Troubleshooting a receiver with a signal tracer. Test signal is provided by signal generator tuned to the receiver's frequency.

In a superheterodyne receiver you sometimes find the converter or oscillator-mixer stages to be at fault. If either the mixer or the mixer portion of the converter is the culprit then the trouble can usually be located using the techniques already discussed. In other cases, though, the oscillator may not be running or may be running way off frequency. If the local oscillator (LO) is a vfo you can generally detect operation using a voltmeter. Measure the small negative voltage on the grid of the oscillator while tuning the dial from one end of the band to the other. If the LO is operating, then the voltage will vary. In crystal oscillators you will also observe a small negative voltage on the grid. If the crystal is in a socket try removing it while monitoring that voltage. It should drop as the crystal is removed. In either type of LO you can use an oscilloscope to actually view the oscillator signal provided that the scope has an adequate vertical bandwidth.

On some occasions you will find the LO is operating but at the wrong frequency. If the error is great enough, the result will be a dead receiver. If you have a frequency counter or other means of measuring frequency then that should tell the story. In other instances, though, you can use a substitute signal from either a signal generator or vfo. In either event the signal must be at the receiver dial plus or minus the intermediate frequency. Even a lowly grid-dipper can be used in some

receivers. If the LO is running on the correct frequency you will note birdies (heterodynes) in the output. If the LO is *not* on frequency expect to hear radio stations!

That elaborate test equipment is *not* needed for most amateur radio troubleshooting jobs may come as a surprise to many. Adding to that belief is the fact that most commercial communications shops have a mighty large investment in multi-kilobuck instrumentation. Even the professional, though, needs only very simple stuff in *most* situations. It is to satisfy FCC regulations and to handle the really difficult service that he needs his equipment. You have one advantage over the pro in that your livelihood doesn't depend on your service work and you can always take your set to him if you fail.

The absolutely most basic piece of test equipment is the multimeter. For most jobs either a simple vom or electronic voltmeter is needed. If a vom is selected make sure it is a model with sensitivity over 20,000 ohms per volt. Between hamfests, auctions, low-cost kits, and the Japanese imports there is no reason why every amateur shouldn't own one. Don't be too awestruck by those modern digital jobs. Your little vom will do your job just as well! Besides, have you ever seen what happens to some of those new digital multimeters when a transmitter is turned on a few yards away? They go bananas!

As described earlier you can obtain a signal tracer by either construction or through a conversion job. Do *not* use any ac/dc dquipment for the conversion, however. The resultant "instrument" may well be lethal. Use only equipment with a power transformer!

Oscilloscopes are often described, quite rightfully, as the most useful piece of electronic test equipment. If, for example, the scope is dc coupled, you can use it in lieu of a voltmeter. If the scope is calibrated, the smallest full-scale range and the overall accuracy is better than most vom/vtvm instruments. Oscilloscopes tend to excite the newcomer and many an auction has seen a frenzy when an old clunker came up for a bid. Frequently you see old, junked, scopes fetch prices all out of proportion to their worth. Because of this a word of

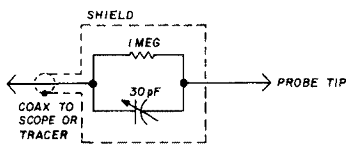


fig. 4. Low-capacitance probe for use with a signal tracer or an oscilloscope. The 30 pF trimmer is adjusted for best squareness of a 1 kHz square wave.

caution is in order. Before you pay $75 to $100 for some elderly 500-kHz, recurrent-sweep model which is old enough to vote, look to the kit manufacturers and some of the imports. For not too much more money you can get a lot more value.

some advice

1. Troubleshooting a dead receiver or other piece of equipment is made easier by one simple fact: it once worked normally! In de-bugging new equipment, on the other hand, you have to give due consideration to that nagging fear that all your troubles may be due to either a false premise in design or a misplaced decimal point. Not-so with repair troubleshooting. Don't wax philosophical and look for deep, subtle troubles. The occult or difficult forms only the tiniest percentage of all amateur rapairs. In other words, think like a repairman, not like an engineer!

2. Don't *ever* use "alignment" as a troubleshooting technique. One of the quickest ways to spot the newcomer or the incompetent old timer is by the speed with which he grabs an alignment tool. Defects which cause a re-

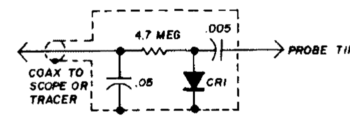


fig. 5. Demodulator probe for use with a signal tracer, an oscilloscope or a voltmeter. Diode CR1 is a germanium type such as the 1N34A, 1N60, etc.

ceiver suddenly to quit seldom, if ever, arise from "alignment." Lock that darn diddle stick up until you know for sure that alignment is indicated. You may find that you never need that tool again! Contrary to some "official" advice to amateurs, receiver alignment does not change enough to justify realignment every year. Although some minor improvement may be affected, alignment just loosens coil slugs and creates other damage that will either keep you off the air now or more often in the future.

3. Avoid bizarre repairs. One such fix which is heard of from time to time is the bending of tuning capacitor plates to make the oscillator track properly. Since it is a fair bet that the alignment of that oscillator was once correct this is a fool's method. I prefer to believe that the engineers who originally designed that receiver and the capacitor in it had a whole lot more sense than I do, at least concerning that receiver. In a situation involving a friend of mine I learned that the best procedure is to find out why the local oscillator shifted frequency. There are a lot of reasons for LO mistracking including use of the wrong brand of replacement oscillator tube! In this case the friend called and asked whether he should allow a buddy to bend the main tuning capacitor plates. I told him to retrieve that set, preferably without damage, and come on down to the shop after hours. There we could properly diagnose the defect and, if indicated, realign the oscillator with proper equipment. It turned out that the trouble was a small mica padder capacitor in series with the local oscillator tank coil. A replacement, in this case obtained locally, solved the 50-kHz shift! Incredibly enough, that other fellow didn't even notice that the error in tracking existed only on one band.

Troubleshooting is not such a gruesome procedure as some amateurs want to believe. Reduced to its simplest philosophical form, all defects in equipment which was once working are due either to the existence of an unwanted path for current or the loss of a desired current path. It is the job of the repairer to determine which is the case, and to either open or close the current path as indicated by the symptoms and his logcial deductions.

ham radio

time-out warning indicator for fm repeater users

An inexpensive transceiver-actuated circuit that inhibits repeater-timer override

Many fm operators have, on occasion, timed out a repeater and since this practice is frowned upon, the habitual offender is branded as a "leadfinger." This article describes an inexpensive timer that, when connected to an fm transceiver, prevents the user from timing out repeater stations. Other timer circuits have been described,[1,2,3] but all require manual triggering and reset by a negative pulse.

circuit

The timer consists of a 556 (U1) and two 555 (U2, U3) IC timers (**fig. 1**). The first half of U1 is connected to trigger on a positive step-input voltage.[4] The trigger voltage is sequenced with the push-to-talk microphone switch. (The methods of deriving the trigger voltage are described later.) The output of U1A is differentiated to trigger and reset U1B simultaneously, which is connected as a one-shot. The time delay is determined by R1, C1 to provide a delay equal to 10 seconds less than the repeater timer. Therefore, using a 60-second repeater, the time delay should be 50 seconds and is approximately found from $t(sec) = 1.1\ R1C1$. When U1B is on, its output triggers U2, which is connected as a flip-flop. U2 drives a green LED, which flashes approximately 80 times a minute during this time delay.

When the 50-second time delay is reached U1B goes low and U2 is disabled, simultaneously triggering and resetting U3, which functions as a one-shot for 10 seconds (set by R2 and C2). A red LED is on for this 10-second interval. At the end of 10 seconds, the red LED goes out and the cycle is completed. For other

By **Howard M. Berlin, K3NEZ,** Department of the Army, Aberdeen Proving Ground, MD, and Adjunct Faculty, Department of Elecrical Engineering, University of Delaware, Newark, Delaware

choices of delay times, the RC combinations can be determined from the equation. The status of the LEDs is shown in fig. 2.

power sources

For mobile installations the power supply voltage is taken from the car battery. For fixed station use, any regulated 12-volt supply can be used. I used the circuit described by WB2EAX[5] because of its simplicity and used an spdt switch to permit the use of the car battery voltage when operating mobile.

triggering

Triggering the timer from a transceiver can be done in a number of ways. Using my TR-22C, I noticed that the microphone switch keys a transistor whose collector voltage is zero on receive and +12 volts on transmit (Q11). In this way, the push-to-talk action controls both the transceiver and the timer. Connection between the timer and the transceiver was by RG-174 coax cable.

If you're hesitant about going into your transceiver, the timer can be actuated by the transmitter rf output signal, since its voltage amplitude is constant. Fig. 3 is a simple circuit for rectifying the rf signal. A suitable resistor may have to be placed in series with the input for transmitters with outputs greater than 10 watts. A quick estimation of the received rf voltage from your transceiver can be made by using Ohm's law for the power absorbed by a 50-ohm load: $V = \sqrt{50P}$ where P is the transmitter output power in watts.

operation

When the push-to-talk switch is pressed, the green LED will flash repeatedly and the red LED will be off. If a single transmission exceeds the first time delay (50 seconds), the green LED will stop flashing and the red LED will light. At this point you have 10 seconds to stop transmitting or else you will time out the repeater, and *all* the LEDs will be off. If the transmission time is less than that of the repeater timer, the timer indicator can be recycled when the PTT switch is again pressed.

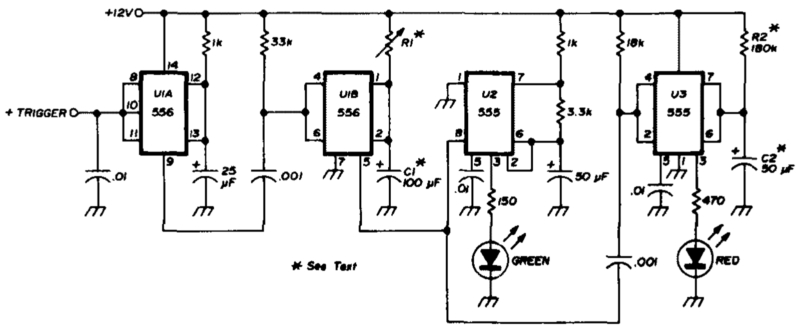


fig. 1. Schematic of time-out indicator.

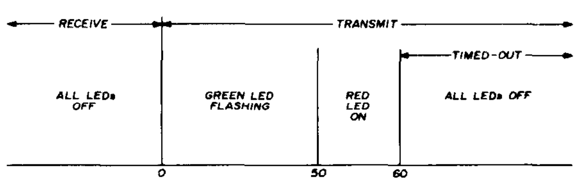


fig. 2. Status of LEDs for a transmission with a 60-second repeater.

other comments

Some of the more astute readers may wonder, "Why go through the trouble of triggering the IC timer by a positive voltage when the 555 and 556 ICs are normally triggered by negative pulses?" Well, I had a few of these chips around and wanted to experiment by wiring them in different configurations. Otherwise, a simple transistor inverter switch with a differentiated output will work as well in place of U1A, and a 555 could be used for U1B with appropriate pin connections. A 741 op-amp could also be used but requires a positive and negative supply. If negative triggering is preferred using the voltage from the transceiver circuitry, U1A can be eliminated.

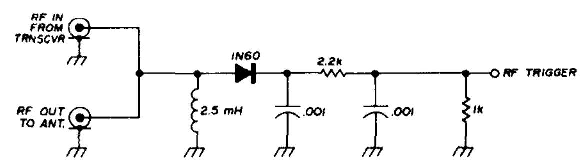


fig. 3. Circuit to trigger the timer from an rf source.

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3. H. Vordenbaum, W5ZHV, "Automatic Reset Timer," *ham radio*, October, 1974, page 50.
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ham radio

MICROPROCESSORS

microcomputer interfacing: accumulator I/O versus memory I/O

When data is transmitted between a microcomputer and an input/output device, three actions must occur simultaneously:

1. The microcomputer must select the specific input/output device that will either receive or transmit eight bits of data.

2. The microcomputer must indicate to the specific input/output device when the bidirectional data bus is available for data transmission.

3. The data must be transmitted between the microcomputer and the input/output device in a very short period of time, typically of the order of microseconds.

In a preceding column,[1] we discussed *accumulator I/O,* in which data is exchanged between the accumulator and an external I/O device. A significant disadvantage is associated with such an interfacing technique: only a single origin or destination for data exists. A typical microprocessor chip, such as the Intel 8080, has in addition to the accumulator a variety of internal general-purpose registers that can exchange information with memory. These registers include the B, C, D, E, H, and L registers, each of which is an 8-bit register. From a programming standpoint, it would be very useful to be able to exchange data between any of these registers and any external I/O device. This is the subject of this month's column.

If you desire to exchange data between a general-purpose register and an external I/O device, you would employ an exciting interfacing technique called *memory I/O* or *memory mapped I/O.* The basic gimmick behind this technique is quite simple: *you treat the input/output device as if it were one or more memory locations.* By doing so, you have the opportunity of employing microcomputer instructions such as MOV, STAX, LDAX, SHLD, LHLD, STA and LDA in the 8080 microprocessor instruction set. These instructions transfer data between registers and memory locations.

The differences between accumulator I/O and memory I/O is best understood with the aid of a specific examply of an interface between an 8080-based microcomputer and an external I/O "device." In this case, the "device" is the Intel 8255 programmable peripheral interface (PPI) IC. This chip has 24 I/O pins, shown as PA, PB and PC in **fig. 1** and **fig. 2**, that can be wired directly to any digital device having TTL-compatible logic.

An 8255 chip appears to an 8-bit microcomputer as either four different external I/O devices or four different memory locations. Four 8-bit registers in the chip can be addressed by the microcomputer:

By David G. Larsen, WB4HYJ, Peter R. Rony, and Jonathan Titus

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

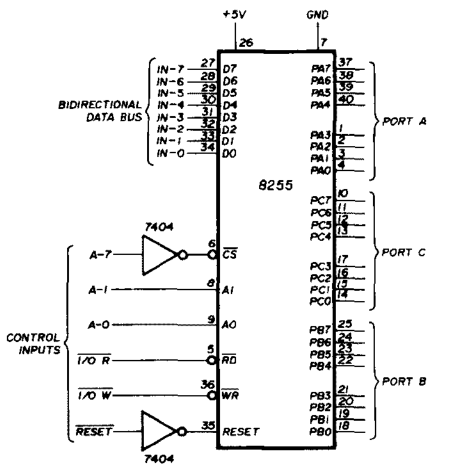

fig. 1. An example of accumulator I/O interfacing between an 8080 microcomputer and the Intel 8255 programmable peripheral interface (PPI) IC. The chip is treated as four different I/O devices.

1. Port A, an 8-bit port that can be configured as either an input port, output port, or a bidirectional I/O port.

2. Port B, an 8-bit port that can be configured as either an input or output port.

3. Port C, an 8-bit port that can be configured as an input or output port or as a pair of control ports, one for port A and the other for port B.

4. An internal 8-bit control register, which determines the specific I/O configuration of the 8255 chip and which can be altered at any time by the microcomputer.

The 8255 chip contains, in addition to the three I/O ports, an 8-bit bidirectional data bit, D0 through D7, that communicates directly with the 8080 chip; six control inputs, $\overline{CS}$, A1, A0, $\overline{RD}$, $\overline{WR}$, and RESET; and two power inputs.

accumulator I/O

In accumulator I/O, the $\overline{I/O\ R}$ and $\overline{I/O\ W}$ function pulses are used to read from and write into the 8255 chip.* The chip is addressed with the aid of bits A-0 through A-7 (or A-8 through A-15 on the 16-bit memory address bus of the 8080 A1 line to select one of the four different registers within the chip. For example, the following program

323_8	Enable the register for port A and allow it to accept data from the accumulator register
200_8	Device code for port A

writes information from the 8080 chip into port A. The port is treated as an I/O device, and an accumulator I/O instruction, 323_8, is employed. A simple change in the control register will turn port A into an input port and permit the use of the program

333_8	Enable the register for port A and allow it to send data to the accumulator register
200_8	Device code for port A

to read information into the 8080 chip from port A.

memory I/O

In memory I/O, the $\overline{MEMR}$ and $\overline{MEMW}$ memory read/write function pulses are used to exchange data between the internal registers within the 8080 chip and the 8255 registers. The entire 16-bit memory address bus can be used to address the chip. As shown in **fig. 2**, we have employed bit A-15 as the chip select input and bits A0 and A1 as the register select inputs. To output data from register B to port A, the following program is used:

041_8 000_8 200_8	Set the 16-bit memory address pointer register within the 8080 microprocessor chip to the memory address of port A
160_8	Move the contents of register B to port A

Once you have selected port A, you can successively output data from other registers in the 8080 microprocessor chip,

161_8	Move the contents of register C to port A
162_8	Move the contents of register D to port A
163_8	Move the contents of register E to port A
167_8	Move the contents of the accumulator to port A

*The accumulator I/O pulse abbreviations $\overline{I/O\ R}$, $\overline{I/O\ W}$, $\overline{MEMR}$, and $\overline{MEMW}$ are those employed by the Intel Corporation, Santa Clara, California. $\overline{I/O\ R}$ and $\overline{I/O\ W}$ respectively correspond to $\overline{IN}$ and $\overline{OUT}$, which we have employed in previous columns. See "Generating Input/Output Device-Select Pulses," *ham radio*, April, 1976, page 44.

Each additional transfer of data requires only 2 microseconds of execution time, which is quite a bit faster than the 5 microseconds required for successive accumulator I/O data.

conclusion

We have demonstrated that both the accumulator I/O and the memory I/O techniques are applicable to the 8255 chip. The specific application will determine the best I/O technique. In some cases, accumulator I/O is best; in others, memory I/O simplifies programming and speeds the transfer of large quantities of data to or from memory. Some microprocessor chips permit only memory I/O interfacing techniques. Such chips frequently have special memory addressing instructions that speed execution time for memory I/O addressing.

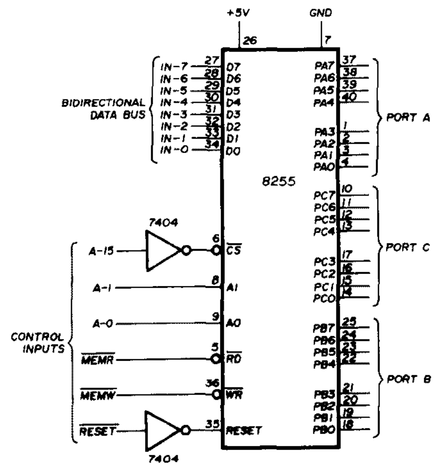

fig. 2. An example of memory I/O interfacing between an 8080 microcomputer and the 8255 chip. In this case the chip is treated as four different memory locations.

The main advantage of the 8255 chip is not in programming or execution time but rather in the ease of wiring of an interface to an external digital I/O device such as an analog-to-digital converter, digital-to-analog converter, digital panel meter, or a digital multimeter. No flip-flops, decoders, or gates are required for the interface; these are all contained within the 8255 chip. In most cases, only SN7404 inverters may be needed to match logic levels between port C, which is usually employed as a control port, and the control pins on the external digital I/O device. Possibly in the future manufacturers will provide I/O interfaces that will permit a digital instrument to be tied directly to a programmable peripheral interface chip.

reference

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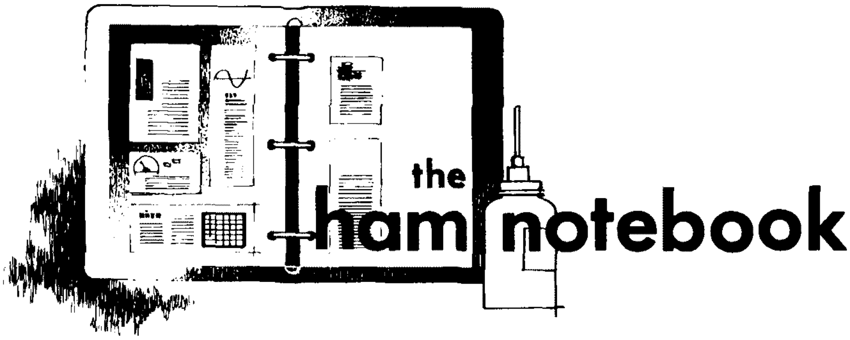


signal-peaking indicator and generator for Regency HR transceivers

The purpose of the peaking indicator is to give a positive indication for maximum received signal strength while pointing a beam antenna or while aligning your receiver. The circuits are simple and are applicable to the Regency HR series transceivers. Since many fm monitor-type receivers use the N5111A or the ULN2111 fm detector IC, the peaking indicator may also be installed in them.

The peaking indicator simply monitors and amplifies the current drawn by the fm detector IC. Under no-signal conditions the current is very small. When a signal is present the current increases proportional to signal strength and can reach several hundred microamperes. The indicator can be a 50 or 100 microampere dc meter. A gain control is necessary to keep the meter from pegging on a strong signal. A multimeter can be used on the low current scale as a workbench aid for aligning receivers.

The actual construction of the peaking indicator is left up to the individual. The circuit is shown in fig. 1 and can be easily installed on the receiver i-f-audio printed circuit. The meter can be mounted in an external case with the gain control. A 0.01 µF bypass capacitor should be connected from the collector of Q1 to ground to prevent negative indication on the peaking meter during transmit due to stray rf rectification. There is an unused pad island adjacent to the 8.2 Vdc track near U102 which can be used.

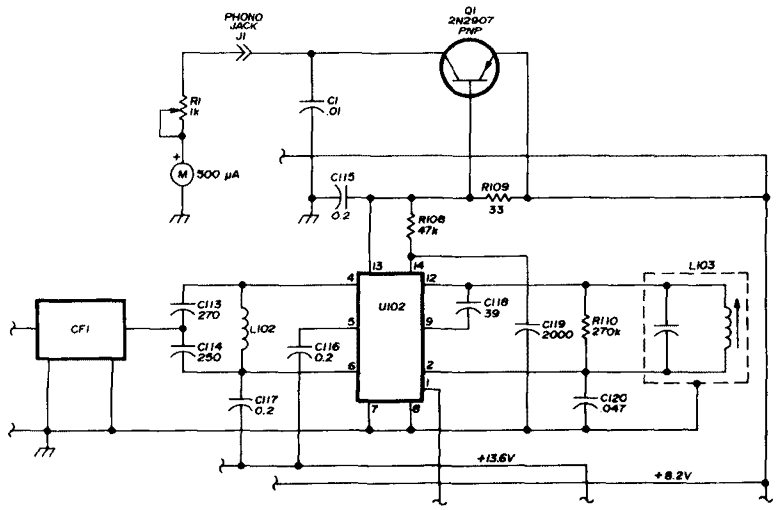


fig. 1. Peaking indicator hookup to Regency fm detector.

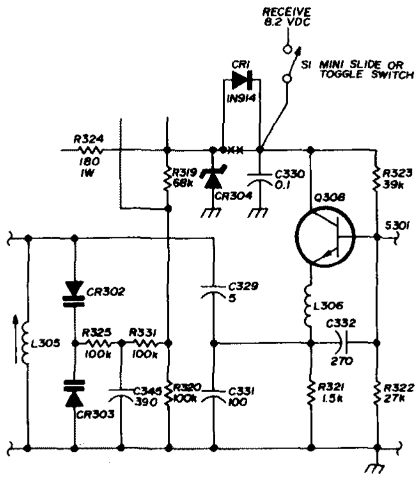


fig. 2. Permanent peaking generator for the Regency HR fm transceiver.

The amplifier transistor and bypass capacitor can be permanently installed on the i-f-audio board with the meter wire connected to a phono jack on the rear of the chassis. The meter may be plugged in as desired when the radio is used as a base station and disconnected when you go mobile.

peaking generator

The purpose of the peaking generator is to provide a stable signal on frequency to peak up a receiver. A simple way to accomplish this is to use the transmitter crystal oscillator right in the radio. This allows selecting and checking each and every transmit crystal in your own receiver.

There are several ways to energize

the transmitter crystal oscillator. The actual implementation is left to the individual.

For a temporary generator the 13.6 Vdc transmit end of R324 can be lifted and connected to a variable 13.6 Vdc source to power the oscillator during the alignment. Varying the voltage changes the signal strength to a certain extent. After the alignment the resistor can be reconnected (see fig. 3).

For a permanent installation the 8.2 Vdc transmit track near the ground end of zener CR304 may be cut and bridged with a diode. Then an externally added switched power source (receive 8.2 Vdc) can be connected to the cathode side of the diode which is the same as the collector of Q308, the transmitter crystal oscillator.

Be sure not to cut the trace feeding R323 which powers the base of Q308. The switch can be mounted on the rear of the chassis with a wire running to the 8.2 Vdc receiver power source. This can be any point common with the circuit trace connected to the orange jumper wire connecting the receiver rf and i-f-audio boards (see fig. 2). This allows switching the transmitter crystal oscillator on without the rest of the transmitter being powered and provides a signal source from the transmit crystal harmonics.

Ray Abraczinskas, W8HVG

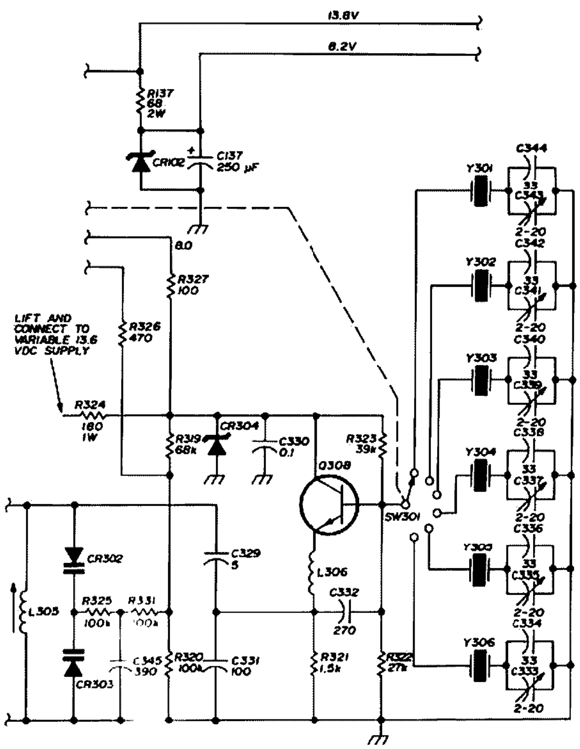
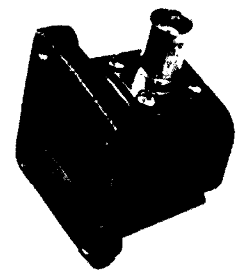


fig. 3. Temporary peaking generator.

marker generator for 3cm amateur band

As more and more radio amateurs venture into the microwave bands, the need for inexpensive but accurate test and calibration equipment increases. One of the most valuable pieces of test equipment is the marker generator. Described below is a simple unit supplying markers every 250 MHz from 8 GHz to above 12 GHz.

The generator consists of a 1N82 diode mounted in a waveguide fixture that is driven with a 300 mW 250-MHz source. The source is a modified transmitter strip from a surplus URT-33 beacon transmitter. Originally the beacon had a positive-ground power supply which is opposite to the rest of



Marker generator provides 250-MHz markers over the frequency range from 8 to 12 GHz.

my equipment. The oscillator and driver stages were supplied from a pulsed six volts whereas the final was supplied a constant eight volts. The only modifications were to strap all the supply leads to a common 6 Vdc supply and isolate the printed-circuit board from the chassis with ceramic insulators and 100-pF capacitors

The beacon originally operated on 234 MHz. Replacing the crystal with a 125 MHz crystal moved the output up to 250 MHz. The slight retuning was

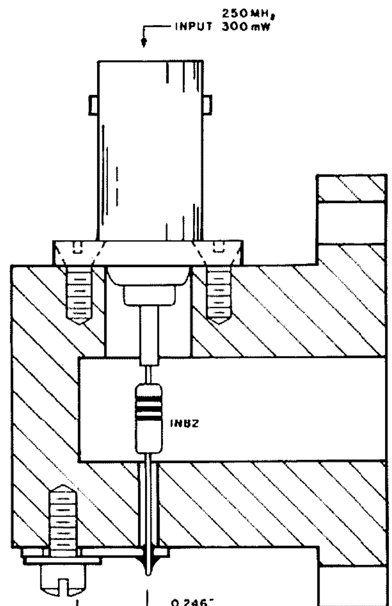


fig. 4. Marking generator for the amateur 3-cm band uses 1N82 diode and modified X-band coax to waveguide adapter.

accomplished by spreading the air-wound tank circuits while monitoring the output power. Incidentally, with +12 volts on the final, I obtained slightly more than 1 watt out which would make a good low power 220-MHz transmitter.

The harmonic generator diode is mounted in a modified X-band coax to waveguide adapter as shown in fig. 4. No bypass capacitors were used because they were found to be unnecessary. The only critical dimension is the spacing between the diode and the end of the waveguide. For best operation this distance should be 0.25 inch ±0.080 inch (6.25mm ±2mm).

The output was examined on a homebrew spectrum analyzer covering 8 Ghz to 12.4 GHz. 20 dB of attenuation was required to prevent the marker generator from overloading the analyzer. The spectrum was rich in harmonics from below 8 GHz to above 12 GHz. The unit provides convenient band-edge markers for the 3-cm amateur band (10.0 to 10.5 GHz) and a marker at the band center.

John M. Franke, WA4WDL

88-mH toroid coils

The surplus 88-mH toroid coils which are available from several advertisers in *ham radio* are often used in audio-frequency filters and other construction projects. Since these coils find widespread use, it's useful to know other values of inductance which are available by simply removing turns from

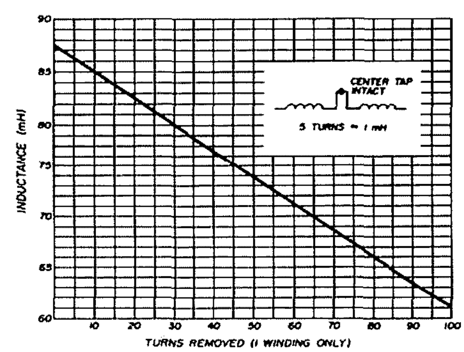


fig. 5. Effect of removing turns from one winding of 88-mH toroidal inductor with center tap intact. Each 5 turns represents approximately 1 mH.

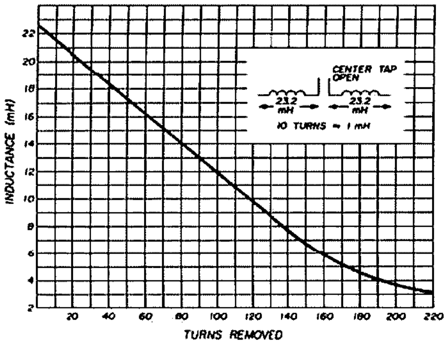


fig. 6. Effect of removing turns from one winding of 88-mH inductor with center tap open. Each 10 turns represents approximately 1 mH.

the core. The coil is manufactured with two separate windings so there are three basic methods for obtaining intermediate inductance values. These are plotted in figs. 5 to 7.

For values between 70 and 88 mH, it is probably easiest to remove turns from only one side. Leave the center tap connected. Use fig. 5, where a reduction of

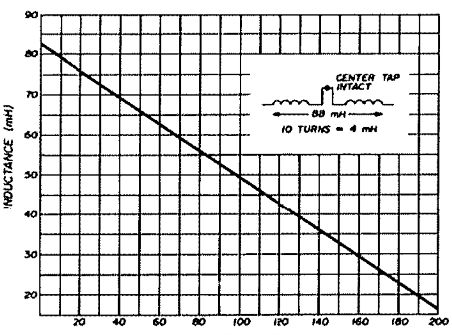


fig. 7. Effect of removing equal number of turns from both windings of 88-mH toroidal inductor with center tap intact. Each 10 turns represents approximately 4 mH.

1 mH is achieved for each 5 turns removed. Fig. 6 gives lower values beginning approximately at 23 mH by opening the center tap and using only one winding. Here 10 turns yield 1 mH.

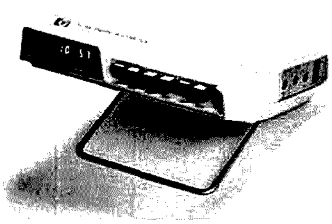
When an equal number of turns are removed from both windings and the center tap remains intact, inductance values shown in fig. 7 are realized.

Measurements were taken with a General Radio Impedance Bridge model 1650A.

Max J. Fuchs, WA1NJG

NEW products

digital multimeter



This new 3-1/2 digit, five function, fully autoranging digital multimeter from Hewlett-Packard sells for only \$225. Voltages measured are from ± 100 microvolts to ± 1000 volts dc and from 300 microvolts to 700 volts rms ac. Resistance measured is from 1 ohm to 11 megohms. Current can be measured from 100 mecroamperes to 1.1 ampere dc and 300 microamperes to 1.1 ampere ac. Autozero, autopolarity and autoranging are built in. The development of fine-line, tantalum nitride resistor technology by Hewlett-Packard has enabled the elimination of more costly discrete precision resistors. Using this new technique, this high-quality compact DMM, with features such as five functions with full autoranging, can be offered at a very low price.

Typical accuracy for dc voltage measurements is 0.5%. Dc current accuracy is 1.0%. On ac voltage ranges, frequency is specified to 10 kHz, while ac current measurement is to 5 kHz. Accuracy of resistance measurements on the three highest ranges is to 0.6% and to 0.4% on the two lower ranges. Open circuit voltage is less than 4 volts. Input resistance on all voltage ranges is 10 megohms with input capacitance of less than 30 picofarads. The 3476 is protected to 1100 volts peak on all ranges. The fuse that protects the ohms function is rated 250 volts rms. The current function is fuse protected to 1.5 amps. No special fuses are required and they can be quickly replaced without disassembling the instrument.

A range hold feature is included that allows the instrument to be locked to any desired range. This feature is necessary when measuring diode resistance. It also makes repetitive measurement faster. The LED readout gives all voltage readings in volts, all resistance readings in kilohms, and all current readings in amperes.

The model 3476A DMM is ac line powered only; model 3467B is ac line powered and also includes rechargeable nickel cadmium batteries. Typical operating time on fully charged batteries is 8 hours. Both units are 2.3 inches (6.5cm) high, 6.6 inches (16.8cm) wide and 8.1 inches (20.6cm) deep. Ruggedness is assured with the high impact resistance polycarbonate case.

U.S. price of the model 3476A is \$225, and \$275 for the 3476B. For more information, contact Inquiries Manager, Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304, or use *check-off* on page 118.

IC test clip

A new product by Continental Specialties Corporation, called Proto-Clip 24, is a welcome third hand for those working with 24-pin integrated circuits. The PC-24 design is patterned after the PC-14 and PC-16. It features a narrow throat, which aids in bringing IC leads up from high-density PC boards. Accidental shorts while testing live circuits are practically eliminated. The PC-24 can be used to inject signals and wire circuits to other boards.

Scope probes and test leads lock onto gripping contact teeth, freeing hands for other work. Contacts are noncorrosive nickel silver, which provides simultaneous wiping action and low-resistance connections to IC leads. Plastic construction eliminates springs and pivots, and a unique molded web ensures thousands of operations. The PC-24 is available off the shelf from local distributors or direct from Continental Specialties Corporation, 44 Kendall Street, P.O. Box 1942, New Haven, Connecticut 06509 or 351 California Street, P.O. Box 7809, San Francisco, California 94104. Write to either address or use *check-off* on page 118.

an introduction to microcomputers

Anyone who has a microcomputer or ever plans to own one should have a copy of this book. It is the nearest thing to a "Bible" for the microprocessor user that I have seen. With books and courses on microprocessors going for \$25 to more than a hundred dollars, this book is also one of the biggest bargains available.

The first two chapters cover the basic vocabulary of microprocessing number systems, and Boolean algebra. Chapters 3 and 4 introduce the components of microprocessor architecture.

Chapter 5 describes logic external to the CPU: memories, I/O, Direct Memory Access (DMA), and system busses. Chapter 6 discusses programming concepts through the description of an imaginary processor instruction set. I prefer this approach to the more usual technique of taking an existing processor, usually the 8080 or 6800, and examining its instruction set. Since every microprocessor reflects the biases of its designers and intended market, a text tied to any existing set must continually digress to explain how other processors differ (or just pretend the others don't exist). Since programming is introduced through examining the instruction set, the reason for, as well as the operation, of each instruction is much clearer to the novice. All addressing modes are described with clear examples, again giving the reader an understanding of why they are provided as well as how they work.

Since the author is a consultant engaged in helping users choose the processor best suited for their application he has an intimate knowledge of most processors presently available. Other authors usually have experience with only a single CPU chip, or are employed by a manufacturer and thus have a vested interest in the success of a particular chip.

For the user who is interested in a specific processor, the last third of the book applies the general concepts to seven commercial CPU chips: The Fairchild F8, the National Semiconductor PACE and SCAMP, the Intel 8080, the Motorola M6800, the Rockwell PPS-8, and the Signetics 2650. Each processor is discussed in terms of its architecture, instruction set (complete instruction

More Details? CHECK—OFF Page 118

sets are given for each processor), I/O and memory structure, and control signals. All support chips (four of them in the case of the 8080) are also described. The level of detail is excellent. For example, there are nineteen pages of information provided for the PACE microprocessor.

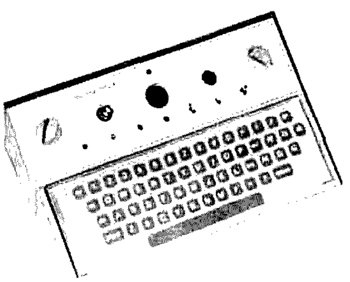
The final chapter is entitled "Selecting a Micrprocessor." The author's contention is that whatever processor requires the minimum external logic to perform a given task is the best processor for that task, regardless of its price, instruction set, or other features.

The beauty of the text is that it can start a chapter by defining what an accumulator is and end up with a discussion of microprogramming without losing the beginner along the way. The more knowledgable reader is assisted in finding the material he needs by a scheme which *outlines* the text in bold-face type, then provides explanatory material and examples in a lighter face. Major topics are further set off by boxed-in keywords in the right-hand margin. The entire design of the book optimizes it as an information retrieval device.

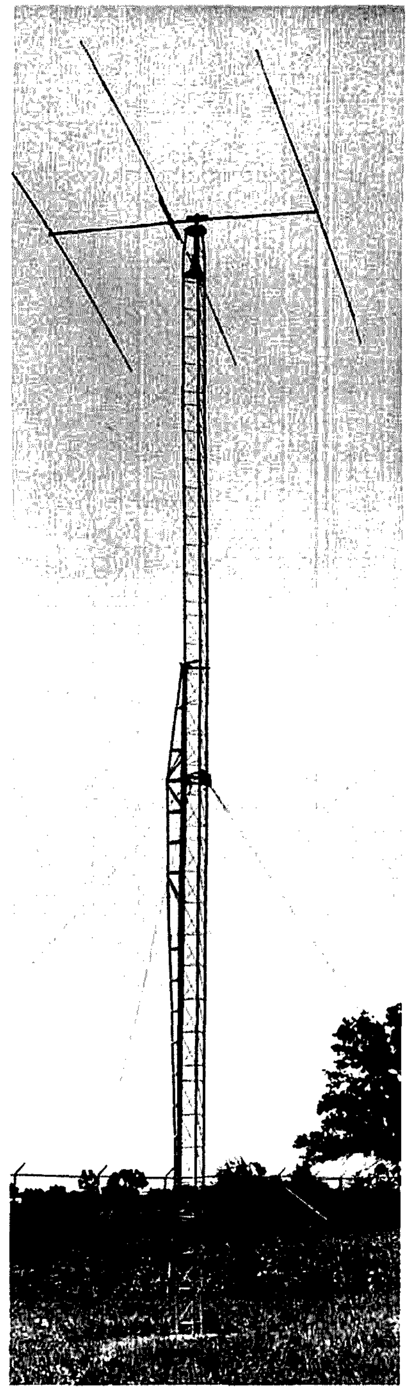
I don't know what more anyone could ask of a general book on microprocessors. Every microcomputer manufacturer should ship a copy of this with each machine he sells — it would stop the complaint that manufacturers never provide enough programming information with their products.

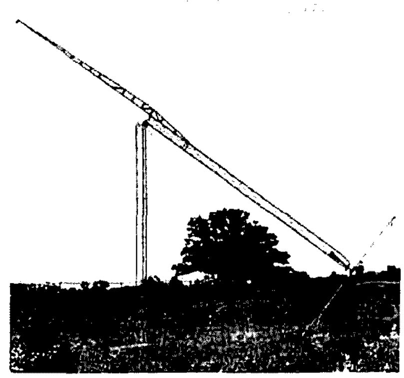
Hard cover, 250 pages, $7.50 from Ham Radio Books, Greenville, New Hampshire 03048.

morsetyper with memory



The Morsetyper model BDC computer terminal can store over six "Quick Fox" sentences or 256 "V" characters in its five registers. It uses this memory





capacity in two modes, QSO or QST, at speeds from 5 to 200 wpm. Both modes allow random speed touch-in while the machine outputs uniformly at any set speed; touch-in slow, output fast, or touch-in fast, output slow. A 56-pad Touch-coder keyboard provides 36 alpha-numerics, 8 punctuation, 7 prosigns, 5 prowords, and error; there is automatic spacing between characters as well as between words. The monitor includes an internal sine wave oscillator with speaker, phones, and volume control.

Distress messages or bulletins can be endlessly repeated, making the machine the solid-state equivalent of a tape recorder but with the advantage of instant reprogramming and lower cost. LED status lights show *store mode, store, full, run message.* The beeper tone indicates *store full* and *monitor.*

The Morsetyper weighs only 10 pounds (4.5 kg) and comes complete with lap mounting holders so you can send from your easy chair. The price is $495. Three other models of the Morsetyper are also available. For more information write Computronics Engineering, 7225 Hollywood Blvd., Hollywood, California 90046 or use *check-off* on page 118.

mobile radio headset

A mobile radio headset that is truly hands-free, operates with virtually no effect from engine and traffic noise, and has no obstruction in front of the face, *Mobil-Ear* has been introduced by JMR Systems Corporation. The heart of the *Mobile-Ear* system is a miniature electric microphone worn on the cheek which picks up the operator's voice through the skin. The technique was developed originally by JMR for use in Navy fighter and helicopter helmets, and for high-noise industrial intercom and mobile systems.

Safety is implicit in *Mobile-Ear's* single-earcup design, which enables the operator to don his headset with one hand, and to hear ambient sounds while still hearing his radio clearly. Comfort is ensured by adjustable eardome pressure and spring-band fit. A unique feature is the system's built-in sidetone, which enables the operator to monitor his own voice.

Two *Mobil-Ear* designs are available.

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Model 10C provides high-quality rf communications (300-3000Hz) for amateur and land-mobile applications with freedom from ambient noise. Model 12 is designed for use in aircraft operations where a single earcup and freedom from ambient noise are required.

In addition to skin contact and acoustic electret microphones and their amplifiers and mountings, other JMR components include mobile radio headsets and helmets. For further information, contact Charles F. Halle, Marketing Manager, JMR Systems Corporation, 168 Lawrence Road, Salem, New Hampshire 03079 or use *check-off* on page 118.

new radio handbook

Since the first *Radio Handbook* was published in 1935, it has grown from a slim 256 pages to the more than 1000 pages of the latest edition, the 20th. A quick check of this new volume, edited by Bill Orr, W6SAI, indicates that about 60 per cent of the text has been revised and more than 115 pages have been added since the last edition was released in 1972.

New chapters in the 20th edition include one on propagation and another on specialized communications techniques. The new propagation chapter starts out with Maxwell's equations, then covers hf and vhf propagation, ionospheric activity cycles, and various vhf propagation modes. The "specialized communications techniques" chapter discusses space communications (Oscar), moonbounce, radioteletype and associated video displays, slow- and fast-scan television, and facsimile.

The chapter on transmission lines has been expanded to include a discussion of wave motion on a transmission line, the use of baluns and matching transformers, and an explanation of the use of the Smith chart. The antenna chapter includes more than twenty new antenna designs, including several high-gain moonbounce arrays for vhf and uhf.

The theory chapters of the new *Radio Handbook* include information on high-frequency broadband amplifiers, vhf solid-state power amplifiers, and the design of vhf circuitry, both for vacuum tubes and solid-state devices. The mathematics chapter has been ex-

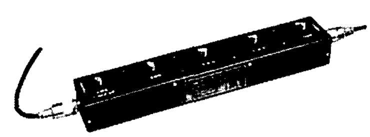
More Details? CHECK—OFF Page 118

tensively revised and includes much new information not generally available to amateurs.

The 20th edition of the *Radio Handbook* also features a number of new projects including an interesting electronic keyer based on a cmos IC which reduces the keyer to a single integrated circuit. This is followed by a discussion of a buffered keyboard for the modern CW operator. For vhf enthusiasts there's a solid-state, ten-watt linear for 432 MHz which requires no intricate stripline construction, and for the advanced builder there's a multi-band communications receiver which features excellent overload and cross-modulation characteristics combined with good sensitivity.

A deluxe 3-1000Z linear amplifier for the DXer or high-power buff is described, as is another linear based on the popular 8877 grounded-grid triode. For the vhf operator there's a new 500-watt amplifier for the 420-450 MHz which uses an 8874 triode. Hardbound, 1080 pages, $19.50 from Ham Radio Books, Greenville, New Hampshire 03048.

TVI filter



The Channel Guard XL-1000, offered by Telco Products Corporation, has features not found in the usual manufactured lowpass TVI filter. It has five sections, each of which is tunable, for maximum rejection of harmonics from transmitters operating below 30 MHz. The XL-1000 handles full legal amateur power with negligible insertion loss.

Used according to directions, the XL-1000 will attenuate harmonics from your transmitter so that no discernible interference from these harmonics will appear on nearby television receivers. It can be inserted between exciter and transmitter or between transmitter and antenna. Delivery is immediate from stock; amateur net price is only $39.95. For more information, write Telco Products Corporation, 44 Seacliff Avenue, Glen Cove, New York 11542, or use *check-off* on page 118.

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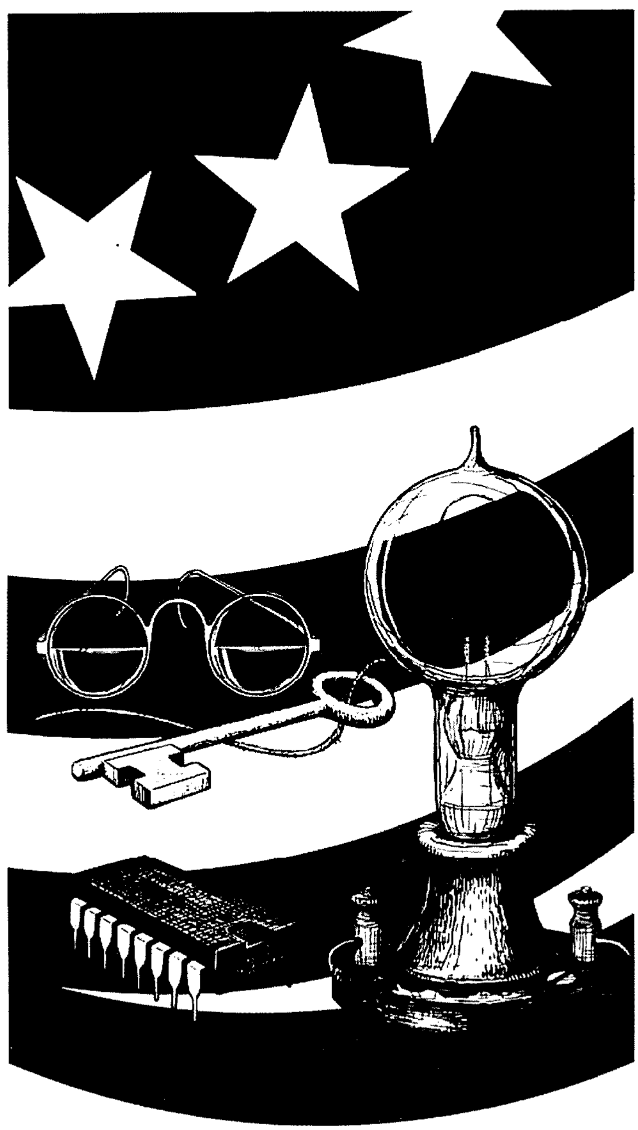
one dollar

ham radio magazine

JULY 1976



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contents

a second look

by Jim Fisk

Although many of the great discoveries in the natural sciences occured in the 18th century, when the Declaration of Independence was signed in Philadelphia 200 years ago, little was known about electricity — most of the great scientific minds of the day were focusing their attention in other areas. However, one of the signers of the document that proclaimed the independence of men as individuals, Ben Franklin, had flown his famous kite some twenty years earlier, thus proving the connection between lightning and electricity.

When Franklin first became interested in electricity in 1746, science — or natural philosophy, as it was then called — was practically nonexistent in America. Americans were forbidden by the British to engage in arts and crafts based on natural phenomena, the Puritan ethic persisted, and men were still cautious about new notions that ran counter to popular belief. Until Franklin's time all that was known about electricity was that when certain substances — such as sulfur or glass — were rubbed, they attracted other light substances, such as bits of paper. No one knew why. Sparks could also be made to jump from the rubbed material to a finger tip, and experimenters noted that the accompanying smell and cracking noise were similar to that produced by lightning. In 1749 Franklin first suggested the "sameness of lightning with electricity," but it was two years later before his paper was published in Paris. The experiment was immediately carried out by two Frenchmen, carefully following Franklin's instructions, one month *before* Franklin's kite flying episode.

Although many others tried the same experiment, not all were as lucky as Franklin. George Richmann, a Swede working in Russia, failed to ground his apparatus - as Franklin had suggested and paid the consequences: a foot-long spark jumped from the rod to Richmann's head and made him the first martyr to the new science.

Although it was the lightning rod that made Franklin a demigod to his contemporaries (the French thought he was the reincarnation of Socrates and slept with his portrait under their pillows), his contributions were much more profound: he unified the disorderly body of existing knowledge that provided a basis for all subsequent advances. Lacking terminology, Franklin invented words as he went along, providing a lexicon of electricity that is still used today. His condenser (or "battery" as he called it) formed an evolutionary link between the short-time sparks of the Leyden jar and the continuous current of the later voltaic cell. He established the positive-negative nature of electricity, hinted at the existance of a basic charge and his single-fluid theory led directly to the concept of electrons moving through conductors. In barely ten years, by trial and error, using simple tools, he had moved a primitive science into the modern world of the 18th century.

After Franklin, the focus of electrical discovery shifted back to Europe, where it would remain for nearly 100 years; Americans were much too involved in the progress of their fast developing country to spend much time or money in nebulous scientific pursuits. It wasn't until 1840, when Samuel B. Morse patented his telegraph, that attention again focused on America.

Morse, a successful portrait painter who knew next to nothing about the basic principles of electricity, had seen some experiments dealing with electromagnetism in Europe in the 1830s and wondered if the effect could be used to send messages over a wire. He made some sketches during his voyage back from Europe, and spent the next three years trying to build the device he had sketched, but nothing came of his work. Lack of knowledge didn't stop him. When Congress offered a $30,000 prize for a 1000-mile system, Morse plunged headlong into the search for a practical telegraph. When one of his colleagues, Leonard Gale, saw one of Morse's unsuccessful machines he pointed out the need for insulation on the windings of the electromagnets, and showed Morse how to arrange the battery circuit. A backer, Stephen Vail, agreed to put up $2000 if Morse would take on his son Alfred. Morse agreed, and it was Alfred Vail who worked out the final form of Morse's code, introduced the key, and reduced the equipment to its final, compact form. It was also Vail who invented the printing telegraph that was patented in Morse's name.

Before 1838, when the patent law was enacted by Congress, only about 500 patents had been granted, but within three years after the patent law more than 10,000 patents were issued. Soon to come were the telephone, the incandescent lamp, the electrical generator, the transatlantic cable and the wireless telegraph. Each of these would lead to thousands of by-products, to major new industries, and to the rapidly advancing electronic technology of the 20th century.

Jim Fisk, W1DTY
editor-in-chief

presstop

MOST NEW AMATEUR HF receivers and transceivers would have to be certified that they meet FCC Part 15 radiation limits under a Notice of Proposed Rule Making recently released by the Commission. Docket 20746 would limit conducted radiation (at the antenna terminals) of receivers capable of tuning in the 26- to 30-MHz region to 100 microvolts from 450 kHz through 25 MHz, with additional radiation limits extending to 3 GHz.
Receiver Radiation Problem came to a head when one popular synthesized CB receiver was found to be radiating sufficient RF in the 37-MHz area that it was interfering with mobile communications in the Power Radio Service.

FCC'S BANDWIDTH DOCKET, 20777, steps on a number of toes and their owners are not at all happy about it. Fast-scan TVers were quickest and most vehement in their reaction to their suggested exile to above 1215 MHz, but the many AM users on 160 meters are also starting to become aware of their jeopardy.
Less Obvious Problems beginning to be discussed are potential conflicts between 850-Hz RTTY, facsimile, slow-scan TV and phone operators on all the present phone bands and the use of modulated CW in the CW segments.

FCC APPROVED ARRL'S TRAINING CONCEPT in mid May, to be tried as an experimental one-year program. The League has proposed a carefully monitored training course of 10 to 12 lessons to be conducted by qualified, certified instructors. Upon satisfactory completion of the course the student would be certified "qualified for Novice license" to the FCC, which would then issue him a license without further exam. By the time you read this approximately 40 clubs and organizations will be teaching courses under this program on a trial basis.
Three Key Stipulations in the FCC's decision favoring the ARRL proposal are an insistence that the integrity of any examinations used in the course be absolute, the instructors must be adequately qualified, and administration of the program must not be restricted to a single organization. Though the FCC is apparently willing to delegate responsibility for determining instructor qualifications to the League, it does not (and probably legally cannot) give the ARRL an exclusive claim on what will surely become a lucrative marketplace.

CHARLOTTE REID RESIGNED from the FCC and will be leaving the Commission on June 30. In the letter of resignation she submitted to President Ford, Commissioner Reid stated that she had just married H. Ashley Barber of Aurora, Illinois and would be returning to Illinois to live later this summer.
Commissioner Reid was also very active in matters affecting Amateur Radio and showed herself to be a real friend of the Amateur service on numerous occasions. She was also an honorary member of AFAR, the Aurora (Illinois) repeater group. Her presence on the Commission will be missed by the Amateur Community.

220-MHZ CLASS-E CB was dealt another blow by a submission filed with the Commission May 3 by the Association of Maximum Service Telecasters, a TV broadcasters' group. Heart of the submission was an April 20 report from A.D. Ring & Associates, a Washington consulting radio engineering firm, stating that severe interference could occur to channel 13 TV reception in an urban area from a 25-watt, 220-MHz mobile at a distance up to 300 feet, while areas where channels 11 and 13 are both in use could experience similar problems out to 1000 feet.

AMATEUR AND CB RADIO RULES have finally been split into separate volumes by the Government Printing Office. First out is the CB volume — Part 95 — which is now available from the GPO or one of its stores in major cities for $1.50 (stock number 004-000-00324-1). Part 97, Rules for the Amateur Radio Service, will become available this summer and also costs $1.50 (stock number 004-000-00325-0). Part 99 for the disaster Communications Service, stock number 004-000-00326-8, is due out momentarily and will cost 75¢. Any or all can be ordered now from the Superintendent of Documents, GPO, Washington D.C. 20402 or the Public Documents Distribution Center, Pueblo, Colorado 81009.

NASA HAS APPROVED AMSAT'S request to "piggy back" OSCAR 8 into orbit on a launch sometime in 1977 or 1978. At this time mid 1977 looks likely, leaving the various contributors to the new Amateur satellite precious little preparation time.
OSCAR 8 Frequencies were proposed at the command station operator's meeting. Tentative choices are: 435.15-435.29 in and 145.850-145.990 out for one mode; 145.850-145.990 in, 435.150-435.290 out for the alternate mode. Beacon frequencies proposed are 435.300, 435.145, 145.995 and 145.845. User comments on all these choices are solicited.

SOUTH AMERICA'S FIRST 432 MOONBOUNCE will be available this summer thanks to Mount Airy VHF Club (the Pack Rats) and the Colombian government. Three Pack Rats, W3HQT, K3BPP and W3HMU, will accompany a complete EME station to Barranquilla, Colombia in time for early August operation. They plan to be active for about two weeks on 432.040 MHz using high power and a portable 16 Yagi array.
Stateside Liaison will be handled through W3KKN and W3NTP at Callbook address or 215 659-3485; HK1BYM will handle the Colombian end. The group plans to field test the complete setup in the June 15-16 VHF QSO party from a portable location.

modern design of frequency synthesizers

A design review of today's synthesizers including a practical circuit for 41-71 MHz that provides low-noise output in 1-kHz steps

All frequency synthesizers use one of two methods to generate output frequencies for use in communications equipment: direct frequency synthesis and synthesis using the phase-lock technique. Direct frequency synthesis has been extensively used in the past and has more or less been responsible for the word "synthesizer," which describes generators that provide accurate and stable frequencies derived from one frequency standard.

This article presents a survey of existing synthesizer technology. The major circuit elements comprising the frequency synthesizer are analyzed, with emphasis on recent design techniques of the phase-locked-loop method to achieve fast switching, low-noise, relatively spurious-free output at high frequencies. Special emphasis has been placed on the analysis of frequency dividers using TTL or cmos logic devices as synchronous counters, as well as phase discriminators using cmos logic. A practical synthesizer circuit for use in the 41 to 71 MHz range is also included, which employs most of the techniques described in the circuit analyses.

direct frequency synthesis

A typical arrangement of this method is shown in fig. 1, in which the desired output frequencies are created by mixing various individual frequencies. The output frequencies are not derived from one oscillator only but are obtained by mixing various frequency components, which are filtered out of a spectrum of frequencies.

Spectral purity. Spectral purity is a basic characteristic that defines synthesizer quality. It's important to distinguish between wide-band performance (that is, performance with sideband noise, which appears symmetrically about the carrier as modulation) and performance with spurious frequencies.

The selective filters used in the direct method determine the spurious-frequency response, and the sideband noise depends on the wideband-mixer power level and crystal-oscillator performance. The signal-to-noise performance of a direct synthesizer, up to 10 kHz off the carrier, is better than that of a free-running LC oscillator. Further off this carrier frequency, LC oscillators are better by definition because these oscillators can't produce wideband noise. Using direct synthesizers, 80 to 100 dB freedom from spurious response is obtainable, and the sideband noise is typically 130 dB/Hz at frequencies more than 20 kHz off the carrier.[1,2]

Switching Speed. Frequency change in direct synthesizers is achieved by switching filters. Switching times with only a few microseconds delay are obtainable; however, phase-coherent switching is not possible.

Disadvantages. The basic disadvantage of a system of this nature is the huge number of components and the requirement of expensive bandpass filters. It is hardly possible to build direct synthesizers above 1 GHz, because the intermediate frequencies will be so much higher that filters having the required performance are not feasible.

frequency analysis

Frequency analysis using phase-locked-loop techniques is a comparatively inexpensive method of obtaining multichannel frequency generators with high stability. For this method the following is important:

The output voltage is determined only from the voltage-controlled oscillator (one), and the exact frequency of this oscillator is determined by the loop. The low-pass filters used in this technique limit the bandwidth of the system where unwanted transients may cause frequency

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

shift. To compare and synchronize the voltage controlled oscillator, which is at a different frequency than that of the reference-frequency oscillator, a mixing and dividing scheme is used for frequencies above 500 MHz, while phase-locked loops using only synchronous counters as dividers are used up to this frequency. This cutoff frequency is determined by the availablity of an integrated-circuit divider. **Fig 2** is a block diagram of such a synthesizer.

Spectral purity. An ideal phase-locked loop, which does not produce its own noise and spurious frequencies, will synchronize and transfer the performance of the crystal oscillator used as a reference to the voltage-controlled oscillator (vco). If the crystal oscillator noise performance is sufficient, the signal-to-noise performance of the vco can be improved, while a noisy reference can degrade the performance of the vco (i.e., an LC or voltage-controlled crystal oscillator). A typical application where the sideband noise and spurious frequencies of an oscillator can be reduced is in the synchronization of microwave oscillators (klystons, microwave tubes, microwave transistor oscillators). A phase-locked loop or synchronizing circuit can improve the performance dramatically. However, there is a limit to the possible improvement because the mixer and phase discriminator will produce noise or spikes.

In addition, the phase-locked-loop circuit acts as an integrating device, and good compensation is possible only when the loop gain is high enough. Because of the lowpass perfomance, close to the cutoff frequency, no noise or spurious frequency improvement can be achieved. **Fig. 3** shows the sideband noise performance of an LC oscillator, a vco synchronized with low loop bandwidth, a vco synchronized with wide loop bandwidth, a standard-quality crystal oscillator, and a high-performance crystal oscillator.

With respect to spurious frequencies, the phase-locked-loop circuit in its pure form (no frequency conversion involved) performs better than direct synthesis because the oscillator does not create any spurious frequencies. Complete freedom from spurious frequencies is impossible, because the loop reference frequency cannot be suppressed to values significantly better than 100 dB.

It is important to understand the following fact: If a frequency is divided, its value is reduced by the factor of the division. At the same time, the amplitude of the fm

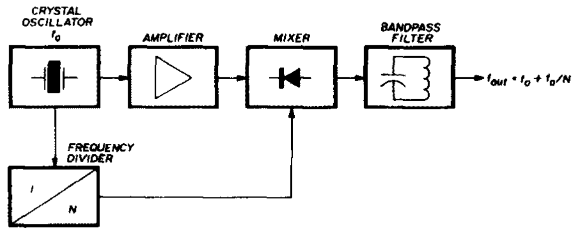


fig. 1. Direct synthesis method in which the output frequency, f_{out}, is equal to the crystal frequency, $f_o + f_o/n$ (division ratio), where n may be changed by using a switch.

spurious frequencies is reduced by the same amount; however, the value of the discrete frequency of the fm spectral line remains the same. For example, two discrete spurious sidebands are located ±100 kHz with respect to a 100-MHz carrier, and the carrier-frequency is divided by 100. The two sidebands are reduced by, say, 40 dB. The modulation index will be reduced by the division ratio, but the modulation frequency will remain constant.

The consequence of this action is that, in the case of a phase-locked-looped system with wide bandwidth, the sideband noise of the reference oscillator will be multiplied up; therefore, the crystal oscillator must be carefully designed in terms of noise performance. This disadvantage can be reduced by using filters of very narrow bandwidth. By doing so, it is possible to build an absolutely spurious-free generator that has the sideband noise response of the LC oscillator (vco). In a circuit such as

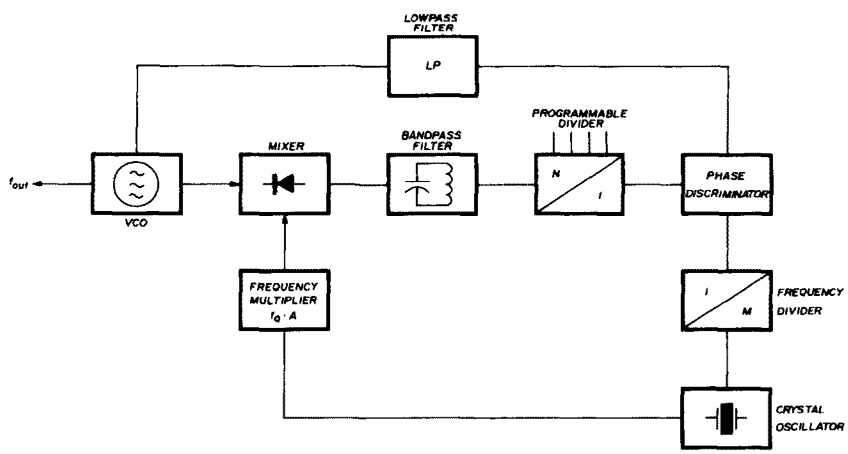


fig. 2. Phase-locked-loop arrangement with a reference frequency oscillator, fQ. The vco frequency is mixed down, using A's harmonics of fQ as an auxiliary frequency to decrease the frequency to a value where integrated circuits are usable. If f_{out} equals 2 GHz and fQ equals 10 MHz, then A might be 180. The bandpass frequency would then be 200 MHz.

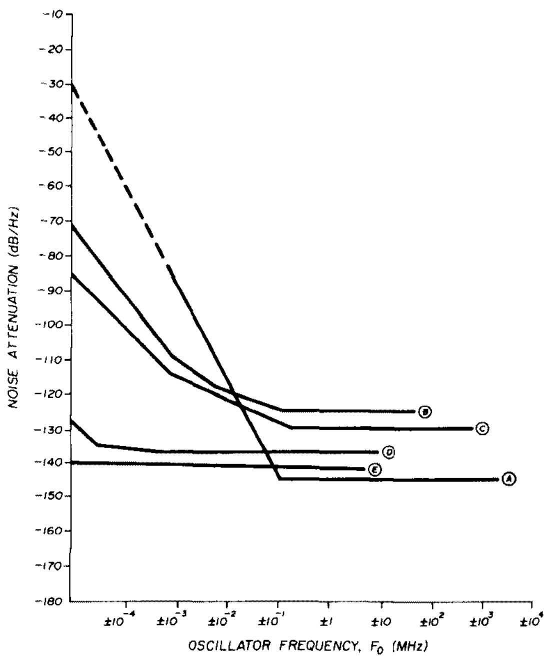


fig. 3. Noise sideband performance or various oscillators. Curve A represents a high-Q LC oscillator; B and C a vco synchronized with narrow and wide loop bandwidth, respectively. D shows the performance of a standard-quality crystal oscillator; E represents a high-performance crystal oscillator.

this, no noise improvement through the phase-locked-loop circuit will be achieved. It is also important to understand that, where extremely low-noise sidebands are required, loop bandwidths between 100 kHz and several MHz must be chosen.

Switching speed. The switching speed is determined by the loop cutoff frequency, depending upon the type of circuit used. In general, switching speed is almost 100 times slower than that of direct synthesizers. An advantage is that phase-coherent frequency switching is possible, which enables digital sweeping. This type of switching is much more accurate and is easily remote-controllable.

Advantages. Phase-locked loop synthesizers have the great advantage that very little filtering is required. In addition, most of the stages can be integrated and very little alignment is required.[3,4] Including harmonic synchronization (sampling techniques), synthesizers up to 10 GHz can be built.

voltage-controlled oscillators

To build low-noise voltage-controlled oscillators, a few design techniques must be considered. Up to 500 MHz, field-effect transistor oscillators show very little noise due to the reduced load they present to the LC circuit. They are superior to bipolar transistors. Since agc is required in some cases, and agc introduces some noise into the system, the performance of some bipolar transistor circuits nearly equals the performance of fets.

In some instances it is desirable to preset the vco to certain frequency bands; eg., 1 MHz wide. This is referred to as "coarse tuning" and is accomplished by using a digital-to-analog converter with sufficient filtering to avoid noise. This technique has the advantage that sample-hold discriminators, which are explained later, can be handled somewhat more easily.

Above 500 MHz only bipolar transistors or gallium-arsenide fets can be used. **Fig. 4** shows a typical field effect transistor vco, and **fig. 5** shows a typical bipolar vco. In some rare cases, voltage-controlled crystal oscillators are used; however, time constants of a few seconds will result, and circuits of this nature are used only where the lock time is of no concern. **Fig. 6** shows such an oscillator.

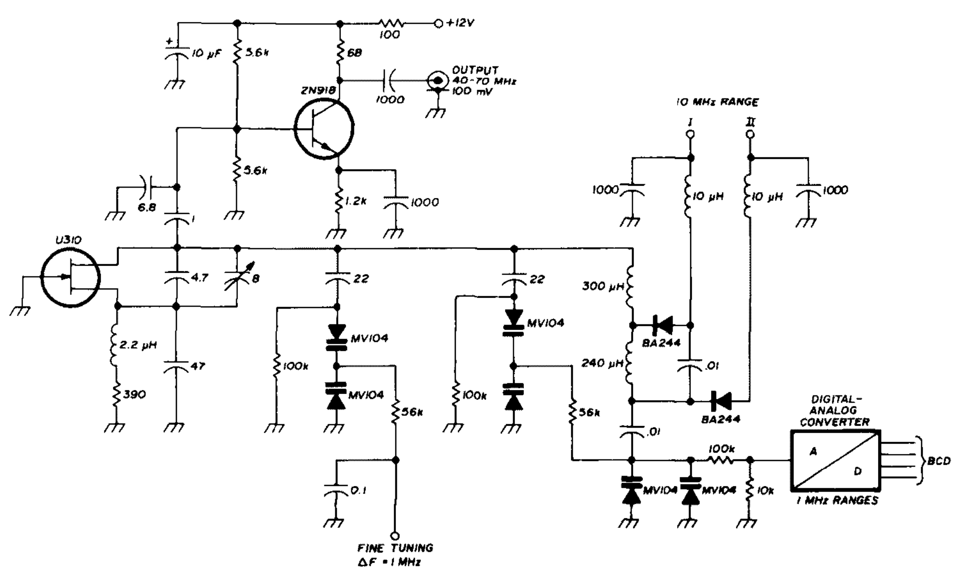


fig. 4. Field-effect transistor vco with coarse and fine tuning, using a digital-to-analog (D/A) converter for presetting.

frequency dividers

TTL or cmos integrated circuits can be used as synchronous counters. Typical ICs are 74192 (TTL), 74C192 and CD4018 (cmos). To extend the frequency range to 500 MHz, so-called "swallow counters" are being used. The most popular swallow counters are the 95H90 made by Fairchild* and the Plessey SP8640. The division ratio of a swallow counter is controlled by two inputs. The counter will divide by 10 when either input is in the high state and by 11 when both inputs are in the low state.

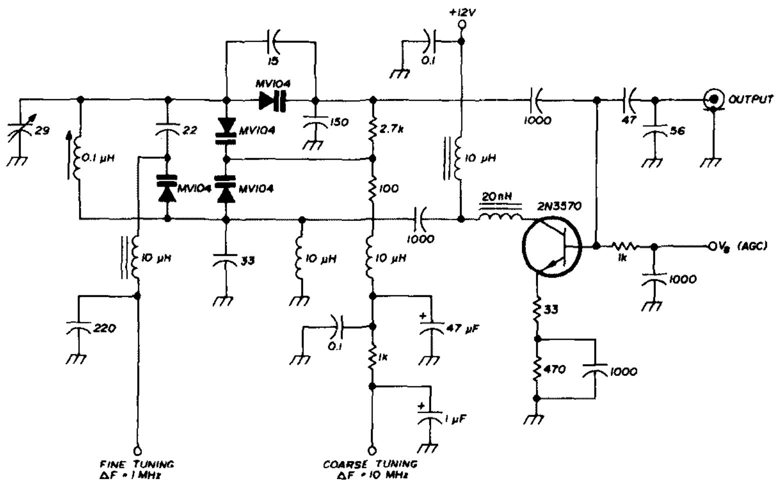


fig. 5. Bipolar vco with fine and coarse tuning.

This 10/11 division ratio enables you to build fully programmable dividers to 500 MHz. The switch counting principle means that high-frequency prescaling occurs without any reduction in comparison frequency. The disadvantage of this technique is that a fully programmable divider is required to control the 10/11 division ratio and that a minimum limit is set on the decision ratio possible — although this is not a serious problem in a practical loop. **Fig. 7** uses a division ratio of *P/P+1*, which is set to 10/11. The **A** counter counts the units, and the **B** counter counts the 10s.

Consider the system shown in **fig. 7**. If the *P/P+1* is a 10/11 divider, the **A** counter counts the units and the **M** counter counts the tens. The mode of operation depends on the type of programmable counter used, but the system might operate as follows. If the number loaded into **A** is greater than zero, then the *P/P+1* divider is set to divide by *P+1* at the start of the cycle. The output from the *P/P+1* divider clocks both **A** and **M**. When **A** is full, it ceases counting and sets the *P/P+1* divider into the *P* mode. Only **M** is then clocked, and when it is full, it resets both **A** and **M** and the cycle repeats.

The divider chain therefore divides by:

$$(M-A)\,P+A\,(P+1) = MP + A \tag{1}$$

*The Fairchild 95H90, which is recommended for operation up to 350 MHz, has recently been superseded by the 11C90 which has a top frequency rating of 520 MHz at room temperature.

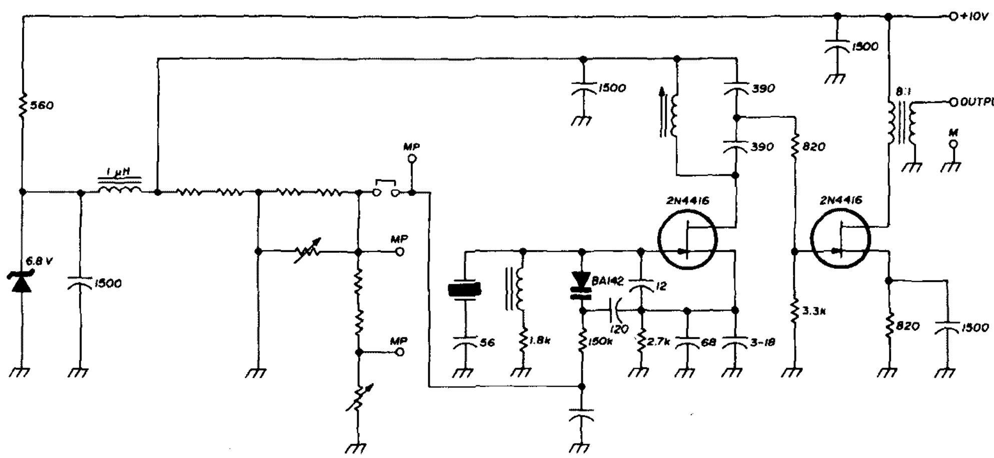


fig. 6. Vcxo circuit with a temperature-compensating network. Instead of using this circuit, a dc voltage may be applied to shift the frequency to desired values. Third- or fifth-overtone crystal oscillators are required for extremely low sideband noise.

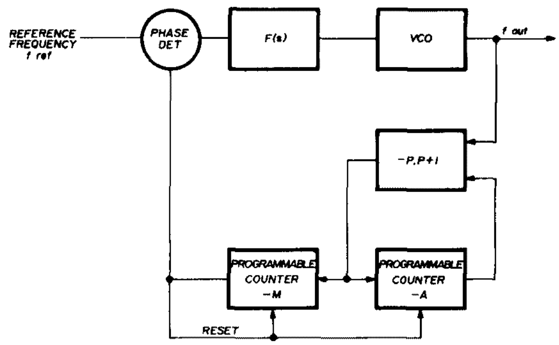


fig. 7. Simplified block diagram of a synthesizer using a programmable prescaler.

therefore

$$f_{out} = (MP + A)\, f_{ref} \tag{2}$$

If A is incremented by one, the output frequency changes by f_{ref}. In other words, the channel spacing is equal to f_{ref}. This is the channel spacing that would be obtained with a fully programmable divider operating at the same frequency as the $P/P{+}1$ divider.

For this system to work, the A counter must fill before the M counter does, otherwise $P/P{+}1$ will remain permanently in the $P{+}1$ mode. There is therefore a minimum system division ratio, M_{min}, below which the $P/P{+}1$ system will not function. To find that minimum ratio, consider the following

The A counter must be capable of counting all numbers up to and including $P-1$ if every division ratio is to be possible, or:

$$A_{max} = P - 1 \tag{3}$$

$$M_{min} = P, \text{ since } M > A \tag{4}$$

The divider chain divides by $MP{+}A$, therefore the minimum system division ratio is:

$$\begin{aligned} M_{min} &= M_{min}\,(P + A_{min}) \\ &= P\,(P + 0) = p^2 \end{aligned} \tag{5}$$

Using a 10/11 ratio, the minimum practical division ratio of this system is 100.

In the system shown in fig. 7, the fully programmable counter, A, must be quite fast. With a 350-MHz clock to the 10/11 divider, only about 23 ns are available for counter A to control the 10/11 divider. For cost reasons

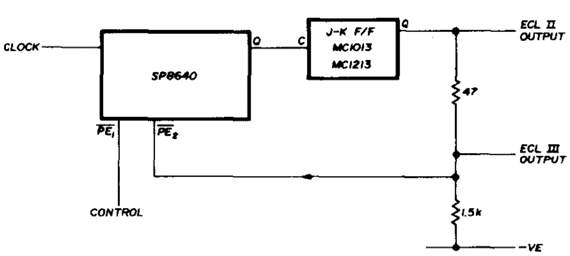


fig. 8. Prescaler with a 20/21 division ratio.

it would be desirable to use a TTL fully-programmable counter, but when the delays through the ECL-to-TTL translators have been taken into account, very little time remains for the fully-programmable counter. The 10/11 function can be extended easily, however, to give a $\div N/N{+}1$ counter with a longer control time for a given input frequency, as shown in figs. 8 and 9. Using the 20/21 system shown in fig. 8, the time available to control 20/21 is typically 87 ns at 200 MHz and 44 ns at 350 MHz. The time available to control the 40/41 (fig. 9) is approximately 180 ns at 200 MHz and 95 ns at 350 MHz.

This frequency division technique can, of course, be extended to give 80/81, which would allow the control to be implemented with cmos, but which would increase the minimum division ratio to 6400 (80^2). This ratio is too large for many synthesizer applications, but it can be reduced to 3200 by making the counter a 80/81/81. Similarly, a 40/41 can be extended to 40/41/42, as

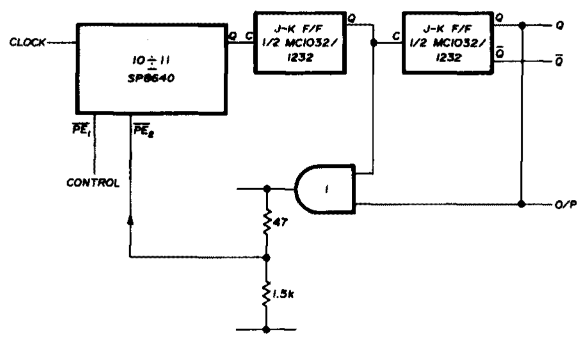


fig. 9. Prescaler system with a 40/41 division ratio.

shown in fig. 10, to reduce the minimum division ratio from 1600 to 800. The time available to control the 40/41/42 is a full 40 clock pulses; i.e., 200 ns with a 200-MHz input clock or 110 ns at 350 MHz. The principle of operation is:

Minimum division ratio

$800 = (20 \times 40) + (0 \times 41) + (0 \times 42)$
$801 = (19 \times 40) + (1 \times 41)$
$802 = (19 \times 40) + (2 \times 42)$

More information can be found in reference 5.

In most cases the oscillator must drive an ECL divider. Fig. 11 shows a simple method using two transistors in a differential amplifier to achieve the nonsaturated voltage swing.

In dealing with counters it must be remembered that, because of the switching action, the counter input represents a high and a low impedance as a function of the status. This means loading the oscillator output stage. Especially when using swallow counters, or so-called variable-modulus counters where unsymmetrical loading occurs, the input signal will show phase modulation. This phase modulation will appear at the counter chain output as excessive sideband noise much larger in magnitude than that contributed by the vco.

To avoid this problem, a low-impedance stage should

be used between the buffer amplifier and the prescaler input; A suitable buffer amplifier is the SN72733, made by Texas Instruments, which has a cutoff frequency of 200 MHz. A common-emitter stage with a 50-ohm load resistor following is a suitable decoupling scheme. Fig 12 shows such an arrangement. Slightly better noise performance is obtained by using the Plessey SP8690 prescaler. This device has a substantially higher input impedance and needs less surrounding circuitry.

lowpass filter is required. Suitable formulas for designing these filters can be found in reference 6. To the best of my knowledge, that book is the best collection of filter tables on the market.

phase discriminators

Various forms of phase discriminators are available. The simplest uses a double-balanced mixer in which two identical frequencies applied to the rf port and local-

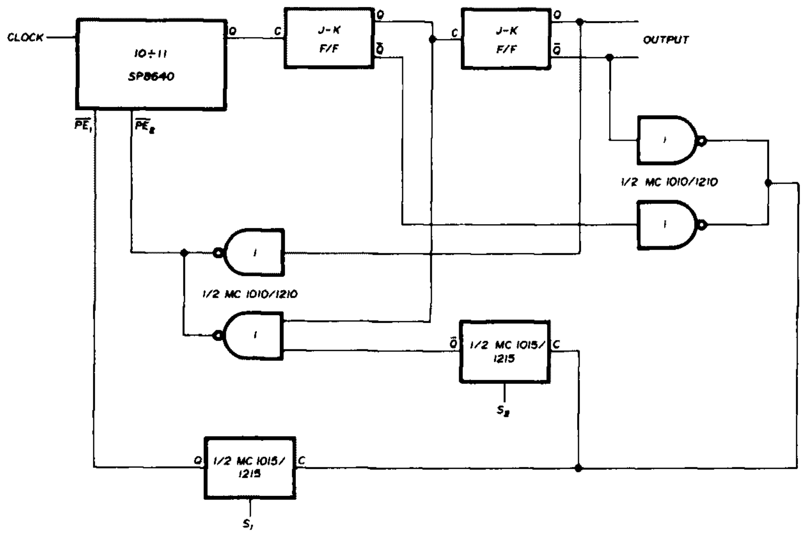


fig. 10. A 40/41 prescaler extended to a 40/41/42 system to reduce the minimum division ratio.

In cases where the synthesizer must be modulated, mixing schemes are used, and the frequency that represents the auxiliary frequency is modulated. Careful selection of the proper mixer and filtering techniques is required to avoid spurious frequencies in the synthesizer. In addition, well-shielded cabinets are required to avoid radiation problems. Fig 13 shows a typical mixing arrangement with adequate filtering. The harmonics of the oscillator frequency to be converted down may produce spurious frequencies, which means that an expensive oscillator port result in a dc output voltage, which must be filtered. Flip-flop discriminators have recently become very popular. The Motorola phase-locked-loop handbook refers to this type of discriminator only because of the ease of its design.[7] However, conventional flip-flop discriminators have significant disadvantages because of the permanent ripple at the output. Therefore, the loop filter cutoff frequency has a tendency of being only 1% or less of the reference frequency so this technique, in practice, does not take advantage of the

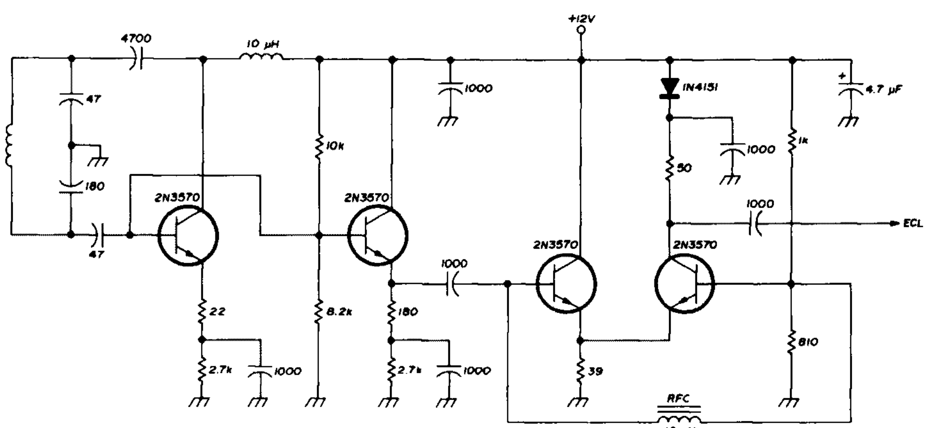


fig. 11. Oscillator circuit with a buffer stage and an ECL voltage translator, using two transistors as a differential amplifier.

possibilities of improving vco noise performance.

Because of the additional introduction of a bipolar transistor as a current charge pump, the sideband noise performance is almost degraded, so the flip-flop technique is seldom used in high-performance synthesizers. Complete information on flip-flop discriminators is given in the Motorola handbook previously mentioned.[7]

The introduction of the RCA CD4046A integrated circuit containing a new type of phase comparator represents a big step forward. Fig. 14 shows a schematic of this cos/mos phase comparator, which is an edge-controlled digital memory network. It consists of four flip-flop stages, control gating, and a three-stage output circuit comprising p- and n-type field-effect transistors. When the p-mos and n-mos drivers are on, they pull the output up to V_{DD} or down to V_{SS}, respectively. This type of phase comparator acts only on the positive edges of the signal and comparator-input signals.

The duty cycles of the signal and comparator inputs are not important, since positive transitions control the PLL system. If the signal-input frequency is higher than that of the comparator input, the p-mos output driver will be *on* continuously. If the signal-input frequency is lower than that of the comparator input, the n-mos output driver will be *on* continuously. If the signal and comparator-input frequencies are the same, but the signal input lags the comparator input in phase, the n-mos output driver will be *on* for a time corresponding to the phase difference.

If the signal- and comparator-input frequencies are the same, but the signal input leads the comparator input in phase, the p-mos output driver will be *on* for the time corresponding to the phase difference. The capacitor voltage of the lowpass filter connected to this type of phase comparator must be adjusted until the signal and comparator input are equal in both phase and frequency At this stable operating point, both p- and n-mos output drivers will remain *off*, and thus the phase-comparator output will become an open circuit and will hold the voltage constant on the capacitor of the lowpass filter.

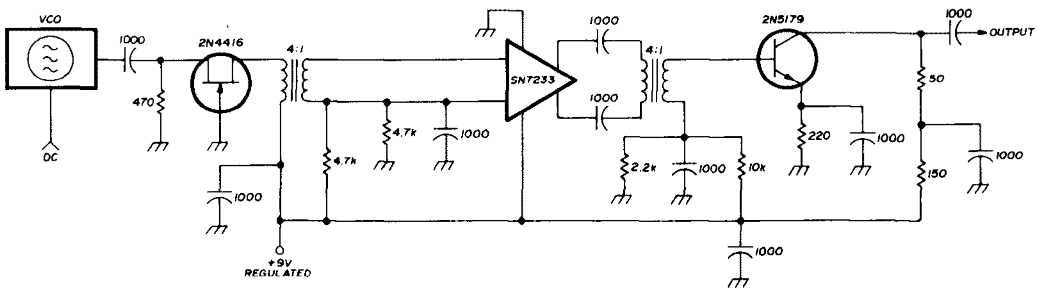


fig. 12. Suitable decoupler arrangement with high gain and low impedance to drive medium-impedance prescalers.

Moreover, the signal at the "phase pulses" output will be at a high level and can be used for indicating a locked condition. Thus, for phase comparator II, no phase difference will exist between signal and comparator input over the full vco frequency range. In addition, the power dissipation due to the lowpass filter will be reduced when this type of phase comparator is used, because both the p- and n-mos output drivers will be *off* for most of the signal-input cycle. It should be noted that the PLL lock range for this type of phase comparator will be equal to the capture range, independent of the lowpass filter. With no signal present at the signal input, the vco is adjusted to its lowest frequency for phase comparator II. Fig. 15 shows typical waveforms for a cos/mos phase-locked loop employing phase comparator II in a locked condition.

Fig. 16 shows the state diagram for phase comparator II; each circle represents a state of the comparator. The number at the top of each circle represents the state of

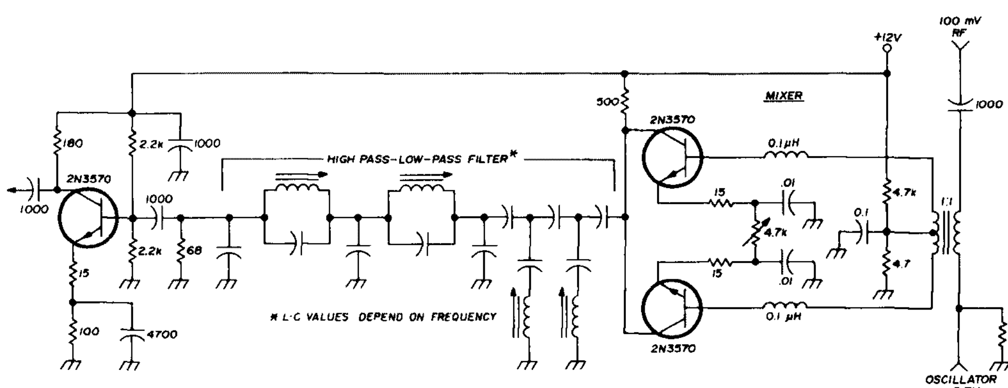


fig. 13. Active mixer for synthesizers requiring 700 mV drive level, together with a highpass-lowpass filter and proper termination.

the comparator, while the logic state of signal and comparator inputs, represented by a 0 or a 1, are given by the left- and right-hand numbers, respectively, at the bottom of each circle.

The transitions from one state to another result from either a logic change on the signal input (1) or the comparator input (C). A positive transition and a negative transition are shown by an arrow pointing up or down, respectively. In the state diagram, it is assumed that only one transition on either the signal input or the comparator input occurs at any instant. States 3, 5, 9, and 11 represent the condition at the output of phase comparator II when the p-mos driver is *on*, while states 2, 4, 10, and 12 determine the condition when the n-mos driver is *on*. States 1, 6, 7, and 8 represent the condition when the output of phase comparator II is in its high-impedance state; i.e., both p- and n- devices are *off*, and the phase-pulses output (terminal 1) is high. The condition at the phase-pulses output for all other states is low.

As an example of how you can use the state diagram shown in **fig. 16**, consider the operation of phase comparator II in the locked condition as shown in **fig. 15.** The waveforms in **fig. 15** are shown in three parts. Part I corresponds to the condition in which the signal input leads the comparator input in phase, while part II corresponds to a finite phase difference. Part III depicts the condition when the comparator input leads the signal input in phase. These three parts correspond to a locked condition for the cos/mos phase-locked loop; i.e., both signal- and comparator-input signals are of the same frequency but differ slightly in phase.

Assume that both the signal inputs begin in the 0 state, and that phase comparator II is initially in its high-impedance output condition (state I), as shown in **figs. 16** and **15**, respectively. The signal input makes a positive transition first, which brings phase comparator II to state 3. State 3 corresponds to the condition of the comparator in which the signal input is a 1, the comparator input is a 0, and the output p-device is *on*. The comparator input goes high next while the signal input is high, bringing the comparator to state 6, a high-impedance output condition. The signal input goes to zero next while the comparator input is high, which corresponds to state 7. The comparator input then goes low, bringing phase comparator II back to state 1.

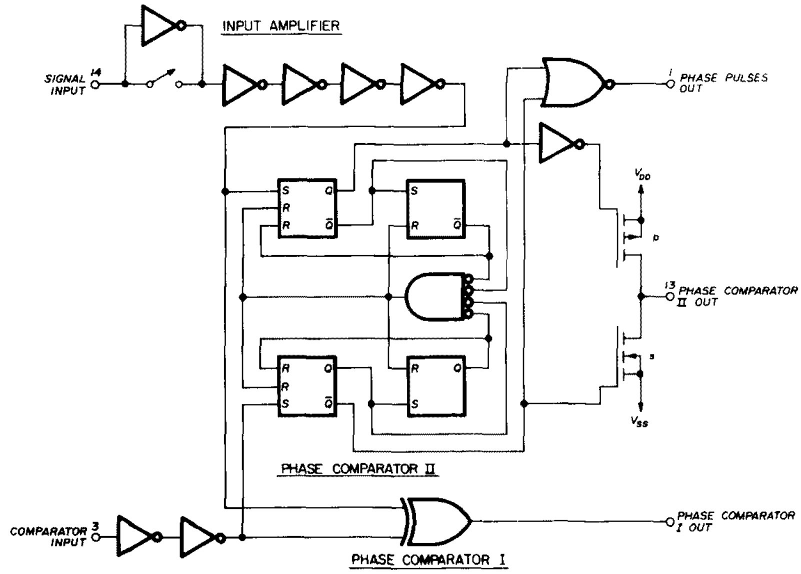
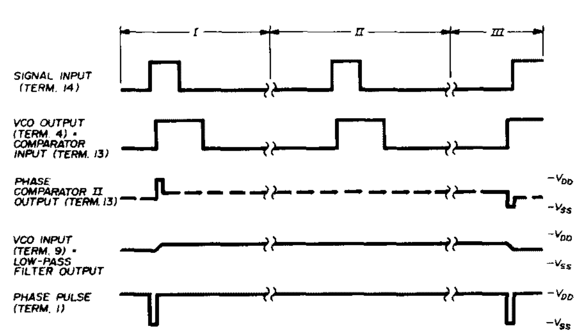


fig. 14. Phase comparator portions of the RCA CD4046A.

As shown for part I of **fig. 15**, the p-device stays on for a time corresponding to the phase difference between the signal input and the comparator input. Starting in state 1 at the beginning of part III, the comparator input goes high first while the signal input is low, bringing the comparator to state 2. Following the example given for part I, the comparator proceeds from state 2 to states 6 and 8, then back to 1. The output of phase comparator II for part II corresponds to the n-device being on for a time equal to the phase difference



NOTE: DASHED LINE IS AN OPEN-CIRCUIT CONDITION.

fig. 15. Typical waveforms for the cmos phase comparator in locked condition.

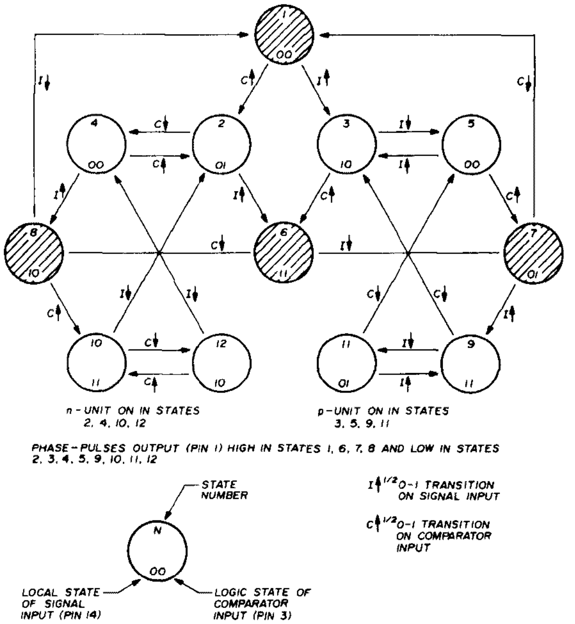
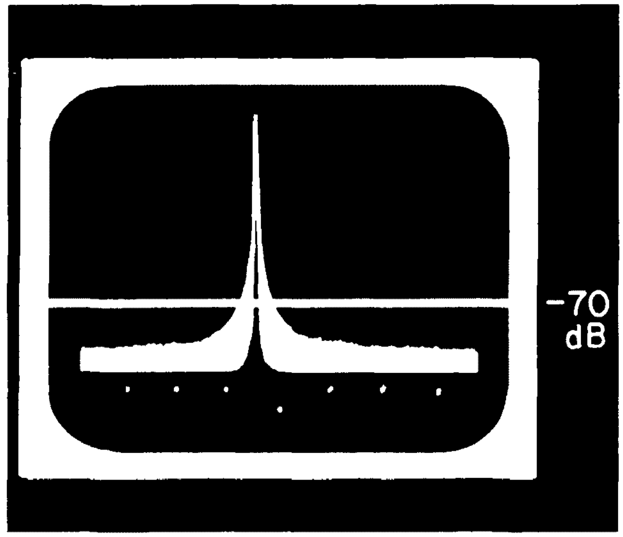


fig. 16. State diagram of phase comparator II in the CD4046 IC.

between the signal and comparator inputs. The state diagram of phase comparator II completely describes all modes of operation of the comparator for any input condition in a phase-locked loop.[8]

As can be seen from fig. 15, the phase comparator II output voltage is a dc voltage, a big advantage over the discriminators used by Motorola. The other important feature is that it uses low-current cmos devices. Building very clean synthesizers can become difficult and expensive because of the shielding involved. Radio-frequency noise created from switching, using TTL devices, will contain much more energy than that from mos devices. In addition, using cmos, the power supply is much simpler. Therefore, it is strongly recommended that cmos be used where possible to take advantage of the inherent good properties of the RCA CD4046 discriminator. This type of circuit comes very close to the sample-hold discriminator, which has been known for quite some time. The sample-hold discriminator requires somewhat more circuitry; however, it offers the best possible reference noise suppression.

The divider-chain output (square wave) is converted into a sawtooth voltage, which is sampled using a switch



Example of a 145-MHz carrier signal. The reference line is 70 dB down; frequency markers are spaced 10 kHz. Excessive noise starts about 60 dB down, which does not fall off completely.

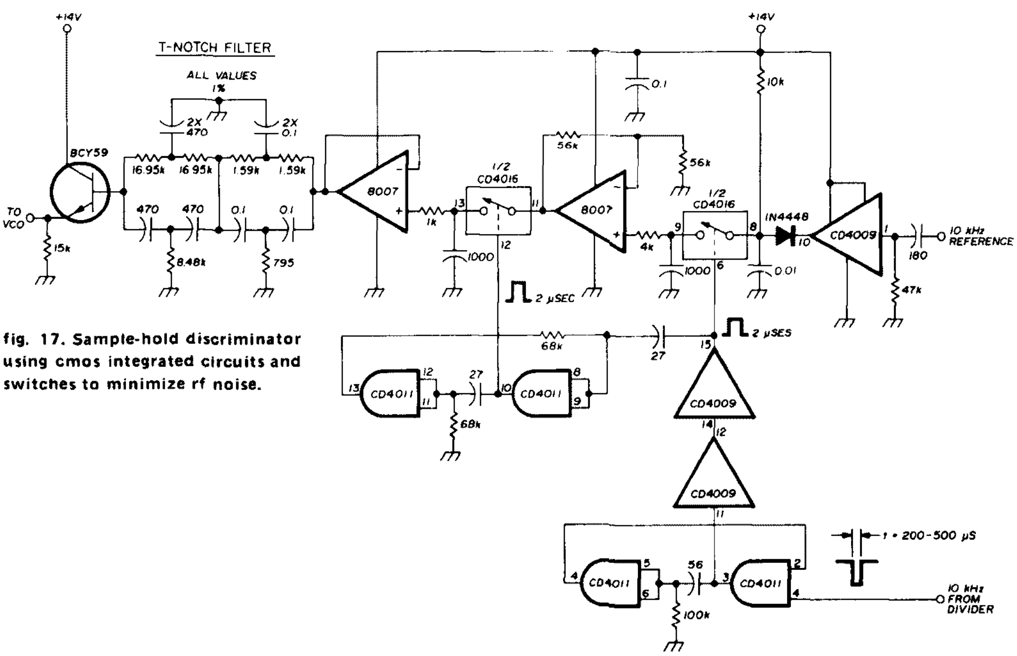


fig. 17. Sample-hold discriminator using cmos integrated circuits and switches to minimize rf noise.

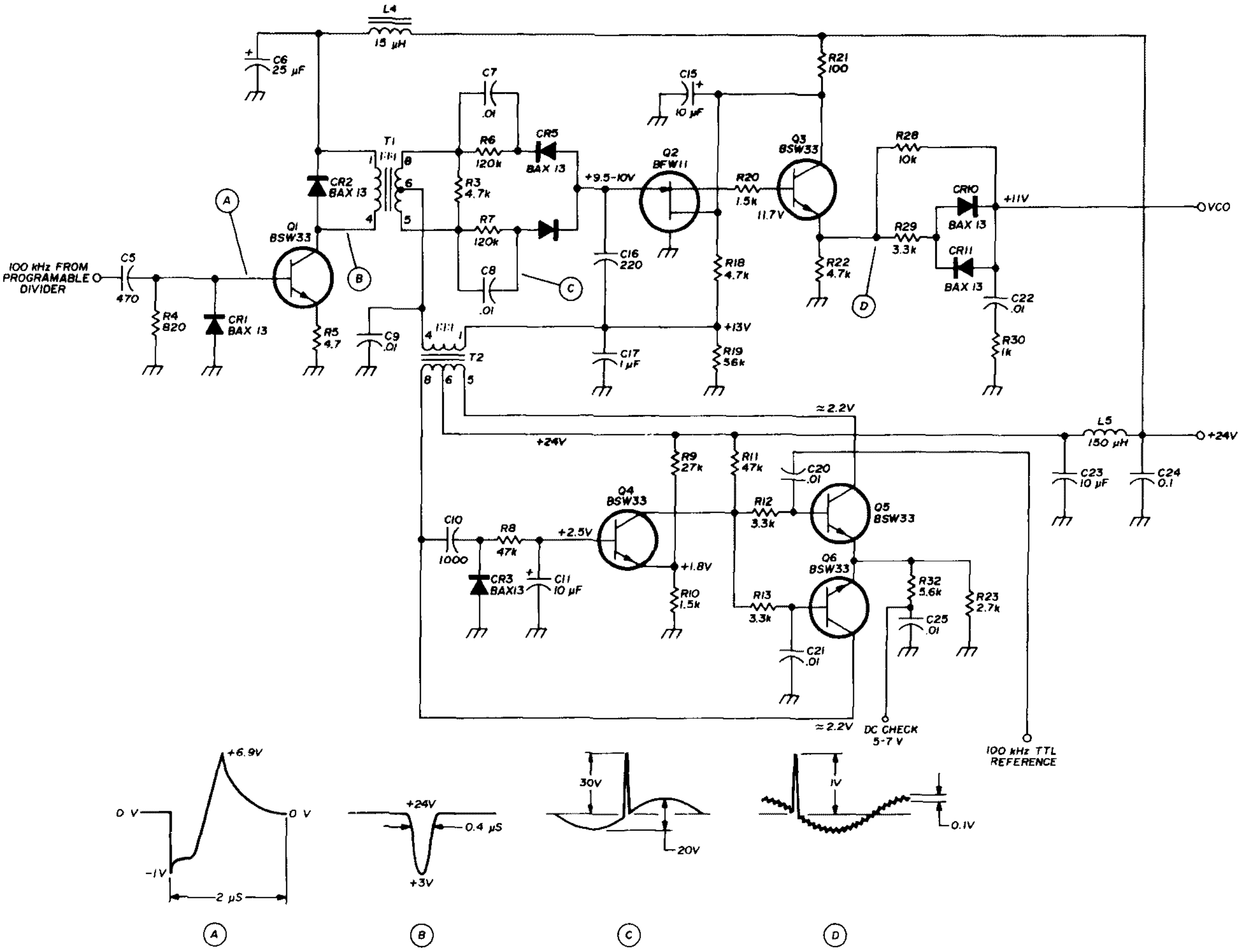


fig. 18. Balanced modulator circuit in a switching arrangement for superior sideband noise performance as a phase comparator.

(sample-hold discriminator). This output voltage is used to charge a capacitor, and the discriminator holds the charge as long as phase and frequency remain the same. If the phase changes, this switch will charge or discharge the capacitor, so that the dc control voltage will also change. Crosstalk or feedthrough is extremely small; the reference noise can be suppressed by 60 dB. However, the fast switching creates spikes, which will produce spurious output.

To reduce spurious output, two sample-hold discriminators are cascaded as in **fig. 17**, which shows a typical arrangement including the waveform-changing stages. The input from the reference divider, which in our arbitrary case is 10 kHz, is used to trigger gate CD4009, and the arrangement with the diode and the RC combination produces a sawtooth waveform (fast charge-slow discharge). The cmos CD4016 receives its input from the reference divider. This input frequency is exactly 10 kHz.

Since the input signal pulse may be a bit too narrow, a one-shot with two gates is used, and the signal now is 2 microseconds wide. The input of the first switch charges a 1000 pF capacitor, and the following Intersil 8007 mos operational amplifier is the low-impedance source for the second CD4016 switch. The input signal to this switch, which is derived from a second one-shot, is delayed; therefore unwanted spikes are suppressed.

The second 8007 operational amplifier drives a T-notch filter. One leg has a 10-kHz resonant frequency; the other 20 kHz. A notch depth of 60 dB can be achieved. The BCY59 transistor is an emitter follower, which drives the vco.

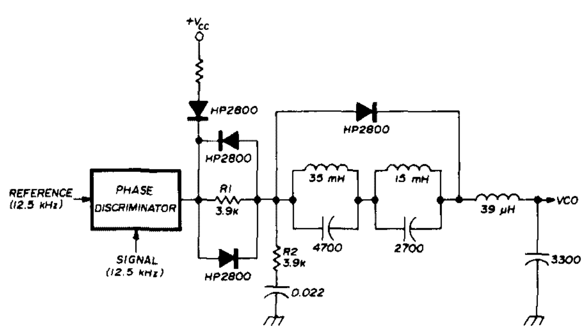
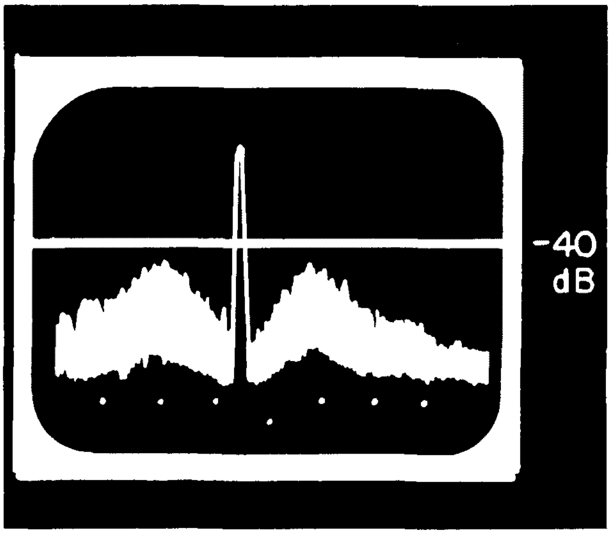


fig. 19. Frequency response filter for synthesizers using a lag filter, two LC tuned circuits as notch filters, and diodes for automatic selection of wider search bandwidth.



Phase-locked loop with a very noisy vco. Loop bandwidth is about 15 kHz within which the sideband noise is improved, as described in the text. Outside the loop bandwidth, noise increases then decreases; however, sideband noise remains at about 90 dB/Hz. Reference line is 40 dB; frequency markers are spaced 10 kHz.

For synthesizers with extremely low sideband noise, an improved version of a circuit similar to a balanced mixer is used. This circuit avoids spikes (in the order of 1 mV in magnitude) which are found in cascaded sample-hold discriminators. To minimize the influence of these spikes, a coarse presetting of the vco is often used, which results in a 20 dB improvement. However, if sideband noise suppression greater than 130 dB/Hz is required, this circuit is still not sufficient.

Fig. 18 shows an arrangement that requires a substantially larger number of components but which provides superior performance. The programmable-divider output (using 100-kHz channel spacing) is differentiated and amplified in transistor Q1. Diode CR2 across the unbalanced-to-balanced transformer provides a voltage similar to a sawtooth waveform. The RC combination R6, C7 and R7, C8 permits the dc voltage to rise to 30

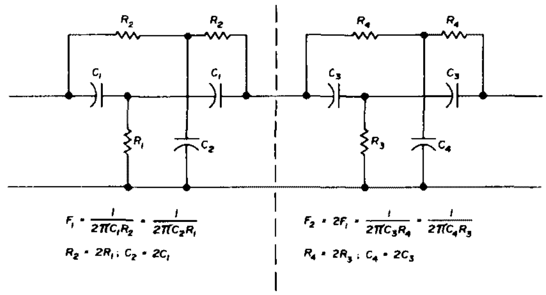


fig. 20. RC twin T-notch filter and the formulas for calculating frequencies.

volts maximum. The 100-kHz TTL reference triggers the one-shot formed by transistors Q4, Q5, and Q6.

The output of this circuit is fed into the center of the bridge of the "balanced mixer," represented by transformer T2. Voltage divider R18, R19 supplies a starting dc voltage, which is brought into the center of this bridge circuit. Transistor Q2 acts as a high-impedance source follower, and transistor Q3 acts as a low-impedance driver for the lag filter. The two back-to-back diodes, CR10 and CR11, are speed-up diodes; their function has been explained earlier. The dc output voltage for the vco contains substantially fewer spikes, and

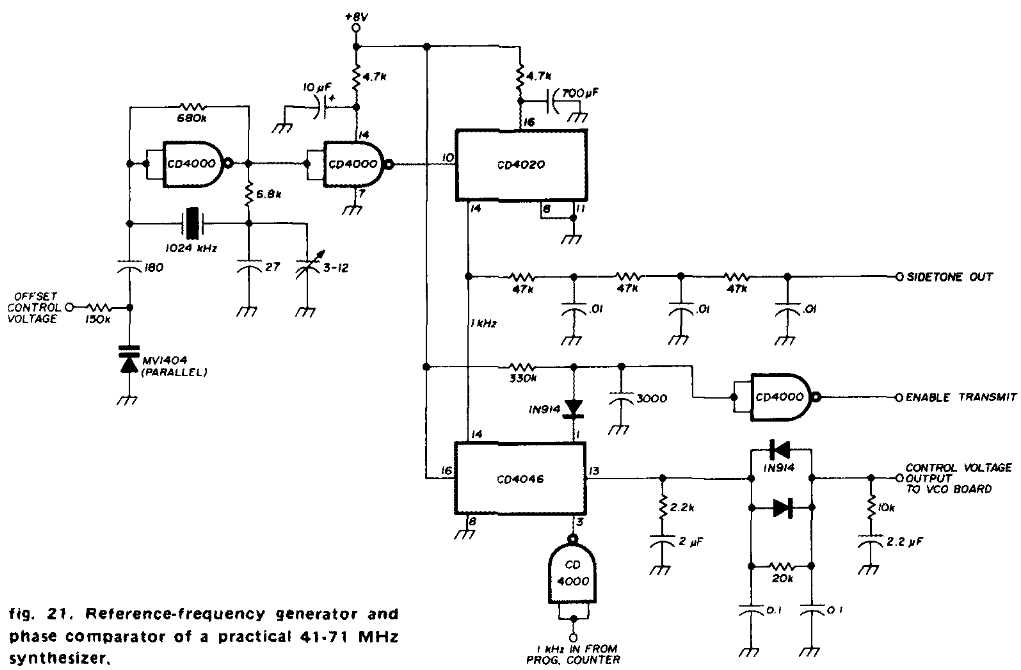


fig. 21. Reference-frequency generator and phase comparator of a practical 41-71 MHz synthesizer.

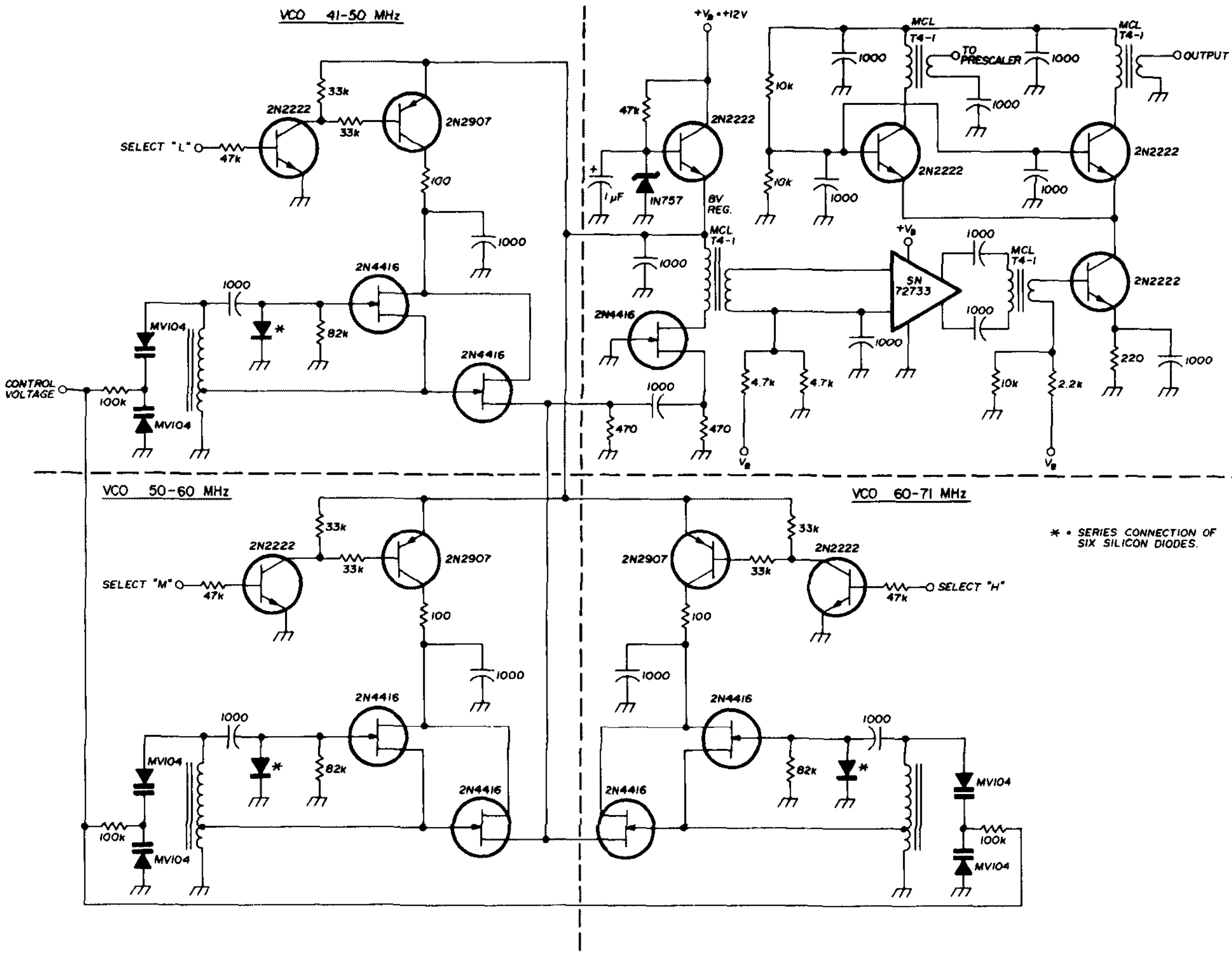


fig. 22. Three independent, low-noise vcos for the 41-71 MHz synthesizer.

the overall loop bandwidth can be made roughly 50% of the reference frequency and still support reference noise suppression. Most high-performance synthesizers on the market use a similar circuit.

response filters

Lowpass filters are required in phase-locked-loop circuits to limit system bandwidth for stable operation. A mathematical treatment of loop stability, as it can be done with the Nyquist diagram, is found in reference 9.

Lag filters are required to set the frequency performance. These lag filters depend upon the type of discriminator used. The Motorola handbook refers to active filter designs where the cutoff frequency is about 1% of the reference frequency. Let's assume that the reference frequency is 10 kHz. The flip-flop discriminator would require a bandwidth of 1% (100 Hz) for 3-dB cutoff. This means that the noise performance of the oscillator can be improved at best between 0 and 10 Hz. All microphonic effects, or sideband noise of the oscillator as such, will remain. In mobile equipment this is most undesirable, because all mechanical resonances will not be compensated.

When analyzing the circuits described in the Motorola handbook, it can be seen that not only are cutoff frequencies down to a few Hz used, but also that the additional transistors used as charge pumps will add noise to the system. Especially the so-called "flicker noise," which appears below 1 kHz, is a very unpleasant effect.

A sample-hold discriminator and the RCA edge-controlled digital memory do not give an output ripple when locked and no shift occurs; therefore, the loop bandwidth can be 20 times wider. In this case, compensation of microphonic effects up to several kHz is available.

As mentioned earlier, in some cases for signal generators, it is only required to synchronize them rather than improve them. The switching speed will then be very slow; e.g., 100 milliseconds. In some cases, as with digital sweeping, this is much too slow. To increase the speed, the filter bandwidth can be increased until the loop has settled. An easy way of doing this is to use two back-to-back diodes across the lag filter. This technique was suggested by Rohde & Schwarz many years ago; however, it is rarely found in the literature.

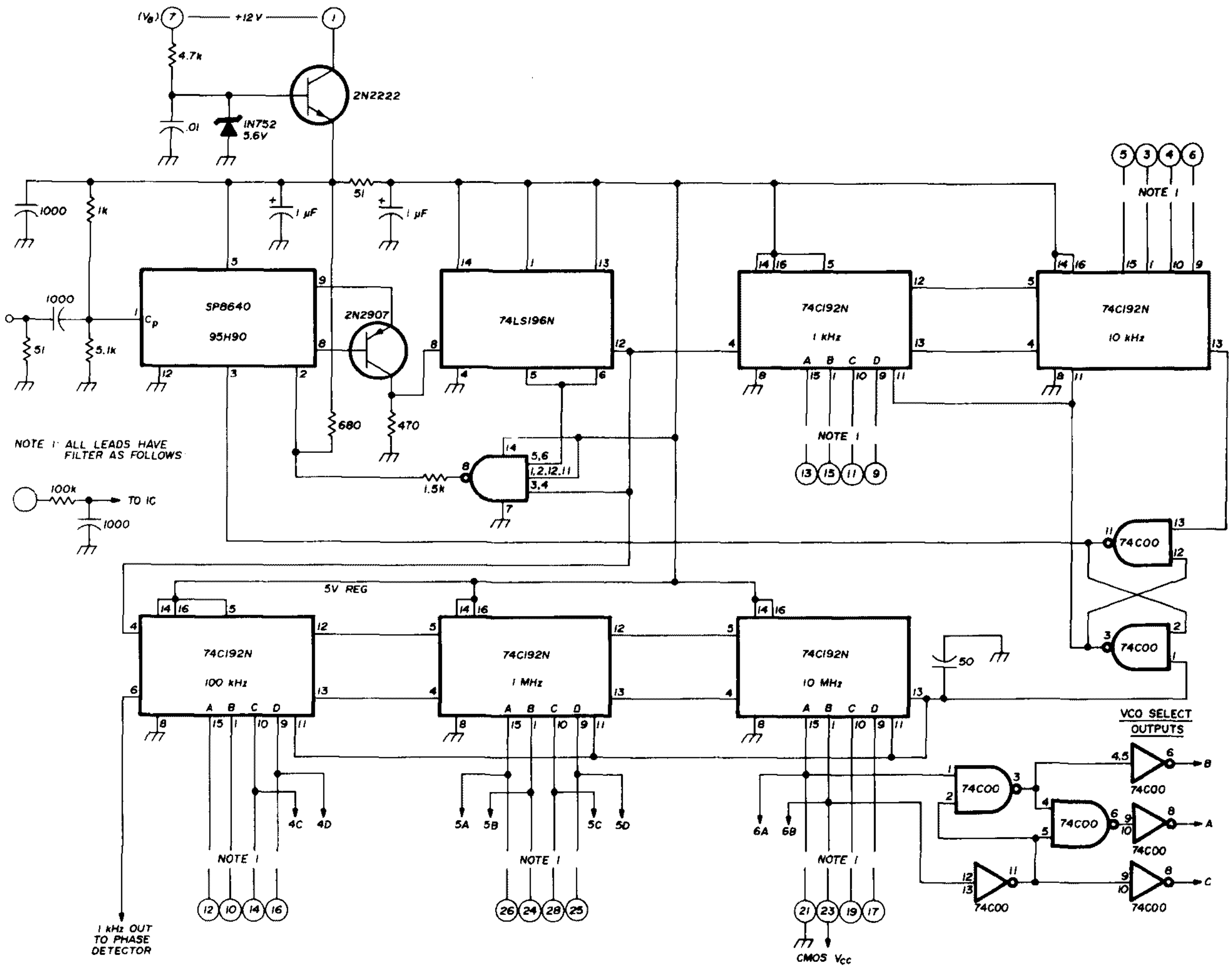


fig. 23. Programmable counter for the 41-71 MHz synthesizer.

If the phase/frequency comparator is in the search mode, the output is a dc voltage with an ac component. This ac component opens both diodes and short circuits resistor R1, which determines the low-frequency time constant. After the loop has settled, the diodes will no longer conduct.

An additional means of suppressing the undesirable reference noise is by using notch filters. **Fig. 19** shows a notch filter with two back-to-back diodes with a series resistor for automatic change of time constants.

A modified Wien-bridge arrangement acting as a T-notch filter can be used to suppress the reference frequency. Together with the RCA integrated circuit or with the sample-hold discriminator, the T-notch filter provides an additional 40 dB suppression. It is highly recommended that two T-notch filters be used, one for the fundamental frequency and one for the first harmonic (10+20 kHz). **Fig. 20** shows such an arrangement. It is also possible to use an LC notch filter for selective suppression, as shown in **fig. 19.**

practical circuit

Figs. **21** through **23** show a complete 41-71 MHz synthesizer that produces steps of 1 kHz. (Using a D/A converter, or better yet, a simple microprocessor, linear resolution can be gained, e.g., the reference frequency can be pulled in such a way that additional resolution, such as 100 Hz, 10 Hz, or 1 Hz, can be achieved).

Fig. 21 shows the reference oscillator and phase detector. A 1024-kHz series-resonant crystal is used, which can be set right on frequency with the variable capacitor. The gate circuit acts as an amplifier to produce an oscillator circuit that is decoupled from the reference divider, CD4020, by a gate. The MV1404 diode can be used for additional crystal-oscilllator frequency pulling.

The RCA CD4020 IC divides the crystal frequency to 1 kHz, and the input is fed to the **CD4046** phase comparator, which was described earlier. In addition, sidetone output is provided and a lock-condition indication is available. The phase-comparator output, which drives the vcos (**fig. 22**), uses a lag filter with back-to-back diodes, as discussed earlier. The 1-kHz signal from the programmable counter is decoupled by a gate and fed to the phase comparator.

Fig. 22 shows the three independent, low-noise vcos

whose outputs are selected by setting appropriate digits on a selector switch. The Texas Instruments SN72733 wideband amplifier is used for decoupling, and a cascade arrangement is used to provide two independent outputs at low impedance.

Fig. 23 shows the programmable counter. The 5 volts required for operation is derived from a simple regulator circuit. The input prescaler uses either the Plessey SP8640 or the Fairchild 95H90, which are pin compatible. The 2N2907 transistor is the ECL-to-TTL transverter. The gate between the 74LS196 and the 95H90, together with the 74C00 at the lower right-hand corner of the schematic, determine the count rate. All other integrated circuits are cmos to keep the power consumption low.

This synthesizer uses most of the techniques described earlier. Because of the back-to-back diodes, the lockup time is 6 milliseconds, and the loop bandwidth after the loop has settled is about 10 Hz. The 1-kHz reference noise is suppressed more than 80 dB and is therefore hardly detectable. This loop bandwidth cannot counteract any microphonics; however, because of its extremely good noise performance and fast switching time this synthesizer is ideally suited for use in a receiver with a first i-f at 41 MHz. Because it's a so-called one-loop synthesizer, it's basically spurious free and the reference noise, as mentioned earlier, is almost totally suppressed.

The total power consumption is in the order of 100 mA at 12 volts dc, and the synthesizer can be built on one PC board, requiring less space than half the size of a normal picture postcard. A very simple circuit, requiring only a few gates, can be built to subtract 41 MHz from the reading; the thumbwheel switches can then be directly calibrated to the receiver input frequency (see fig. 24). In this circuit the upper 74C83 receives the 10-MHz input which must be connected to pins 21 and 23 of the synthesizer input command. The lower 74C38 receives the 1-MHz steps and is connected to pins 24, 25, 26, and 28 of the synthesizer input command.

It has been shown that today's technology permits building fast-switching synthesizers with low-noise and essentially spurious-free output at high frequencies. With new techniques and integrated circuits, these synthesizers consume little power, are physically small, and have high reliability. Especially, if cmos ICs are used, rf noise will be very low and almost no shielding will be required. A shortwave transceiver using a synthesizer based on these techniques will be described in a future article.

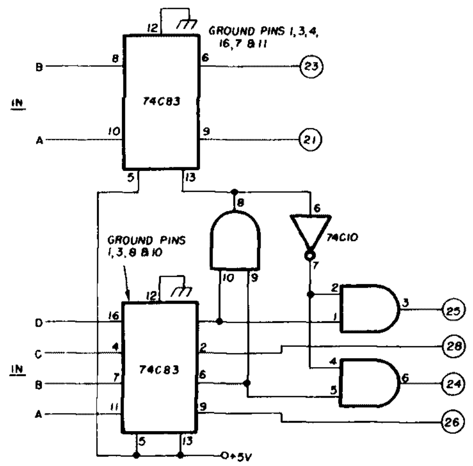
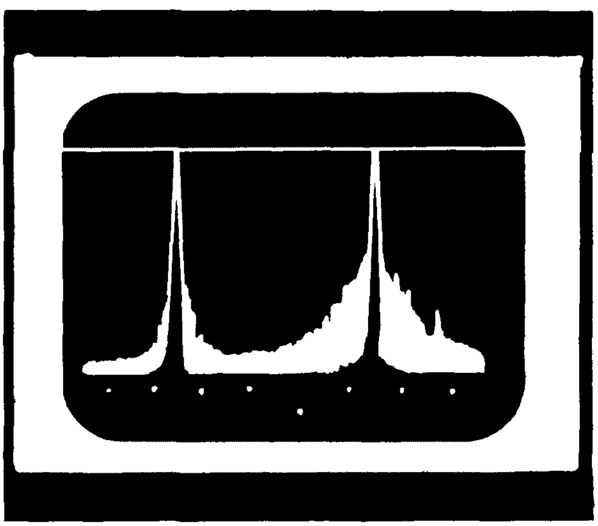


fig. 24. Simple circuit for subtracting 41 MHz from the synthesizer reading so the thumbwheel switches can be directly calibrated to the receiver input frequency.

Comparison of two phase-locked-loop synthesizers with the same reference frequency. Output from balanced-mixer phase detector is shown at left, while the signal at right is from a loop using the technique suggested in the Motorola Phase-Locked-Loop Handbook. All other circuit details are the same. Reference is 0 dB, marker-frequency spacing is 10 kHz, and the carrier is about 150 MHz. Instrumentation used for all photos was the Rohde & Schwarz model EZF/EZFU spectrum analyzer.



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ham radio

wind-generator

characteristics and installation techniques

Windblades and rotors are reviewed and an example is given of an amateur 200-watt electrical generating system

Many experiments have been made over the years with windmills and wind-rotating mechanisms. Several have been adapted as electrical-power generators. The theoretical maximum efficiency in converting wind power to torque at the rotor of a windmill is 59.3 per cent. Many rotating blades and other wind-driven configurations come rather close to this theoretical limit: within 5 to 8 per cent. The present direction of development involves economy in the support structure, gearing, generator, and other accessories. The objective has been to develop a strong, reliable, safe, lightweight, and economical support. Types of blades and rotors are significant in attaining these goals for a specific application.

Air is composed of gas molecules that have mass. Motion of these molecules is called wind. Upon striking a windblade, sail, or similar device, the wind imparts energy of motion. The efficiency of conversion depends on the design of the wind device. The force of such mechanical motion or torque made available at the rotor of a windmill depends on wind velocity, blade size and blade aerodynamics. Power made available varies as the square of the blade radius and the cube of the wind speed. For example, to quadruple the power output, it's necessary to double the blade radius. Doubling the wind speed results in an eight-fold power increase.

The equation that determines wind force in watts that impinges on a slim two-blade propeller is given by*

$$P = 0.005 A V^3 \text{ watts}$$

*When the blade rotational area, A, is in square meters, and wind velocity, V, is in kilometers per hour, the correct formula is:

$$P = 0.0129 A V^3 \text{ watts}$$

By Ed Noll, W3FQJ, Box 75, Chalfont, Pennsylvania 18914

where:

A = Area covered by the blade as it rotates (equivalent to πr^2), square feet
V = Wind velocity, mph
P = Power, watts

Table 1 presents as approximation of the useful power that can be derived from an efficient two-blade windmill in terms of wind velocity in mph (km/hr) and blade diameter in feet (m). An approximate overall efficiency

table 1. Output in watts for efficient two-blader.

blade diameter, ft(m)	wind velocity mph (km/hr) 10 (16)	15(24)	20(32)	25(40)
6 (1.8)	40	140	340	660
12 (3.7)	170	570	1350	2650
15 (4.6)	265	895	2120	4140
20 (6.1)	470	1590	3770	7360

of 30 per cent has been assumed, which includes gearing and generator. Not considered are additional losses in parts of the electrical system such as the charger, battery, distribution system, and inverter (if used). Shown is the relationship between power increases and blade diameter and wind speed.

windmill characteristics

Several characteristics describe the performance of windmills; here they are discussed in terms of two- and three-blade propellers. A simple two-blade arrangement, **fig. 1**, consists of blades, hub, and vane. Two blades are attached to a hub, which is fastened to the windmill rotor. The vane assembly keeps the blades pointed into the wind. To derive maximum benefit from wind power,

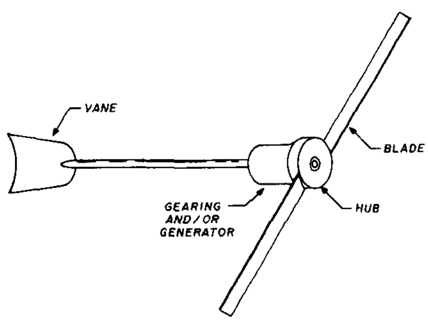


fig. 1. Basic elements of a wind generator.

the deviation from true orientation into the wind should not exceed ±12 degrees.

Another important factor is the blade pitch angle (**fig. 2A**), which refers to the angle of the blade relative to the wind direction. When the relative wind velocity is in line with the blade element, no transfer of wind energy to torque occurs at the axle. For low-speed rotation of a multi-blade windmill, pitch angles of 30° and higher are used. For high-speed rotation, angles are substantially smaller. Small angles are required because the airfoils of many high-speed blades stall in the range of 12° to 14°. However, optimum pitch angle depends on application, desired operating conditions, type of

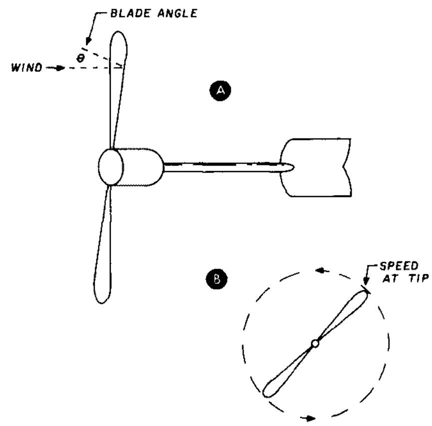


fig. 2. Pitch angle (A) and tip speed (B) of a wind-driven propeller.

blade, preferred angle of attack for blade air-foils, and the use of the wind-rotating system based on wind-speed limits at the site.

Windmills are sometimes classed as either low or high speed. In general a multiblade windmill rotates at low speed and is usually a heavy affair that develops high torque even in a light wind. It is widely used to convert wind energy to some sort of mechanical action such as running a water pump The high-speed windmill employs as few as two blades. It is lightweight and more adaptable to converting wind energy to electricity. Its high rotation speed is adaptable to low-ratio gearing of electrical generators.

Another factor of concern is tip-speed ratio. The tip-speed ratio, **fig. 2B**, is the ratio of wind velocity to tip rotational velocity. It is, in a practical sense, the ratio of wind speed to the speed of motion of the very tip of the blade. Tip speed is often stated as a whole number that compares the blade-tip velocity with the wind

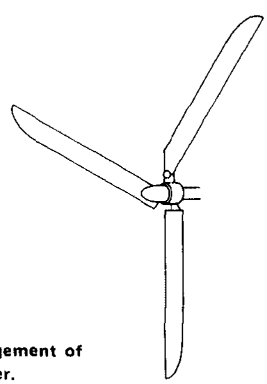


fig. 3. General arrangement of a three-blade propeller.

speed. A ratio of 4 indicates the propeller tip has a velocity four times faster than the wind speed. For electrical power generation, tip speed ratios under 4 are not recommended.

Lift-to-drag (*L/D*) ratio, another characteristic of concern, is an indication of how well the blade is turned by the wind relative to the torque or opposition offered by the propeller to being set into motion by a light wind. This ratio is related to blade construction, size and airfoil. Airfoil refers to the geometric shape of the blade. As in aircraft design, airfoil has a great influence on how well the blade can be turned by the impinging wind. A high-lift airfoil can increase the power output but also increases drag. Nonetheless, a high *L/D* ratio does permit higher output at a lower wind speed. Compromises must be made in establishing the preferred *L/D* ratio in terms of desired power output, weight, and wind-speed range over which the assembly is to operate at high efficiency.

For many low-powered applications a two-blade propeller is effective and efficient. However, in terms of weight and blade diameter, there is a practical limit to the size of a two-blade propeller. A three-blade propeller, **fig. 3**, provides additional power output as compared with a two-blade propeller and reduces periods of vibration with changes in wind direction as well. When the windmill orientation follows the tail vane, the resistance to orientation shift made by a two-blade propeller is in accordance with its position. When in a horizontal position this resistance is maximum. The net result is a jerking movement of the windmill as it follows a wind direction change. This action produces an undesirable stress when the blade is heavy or too large in diameter. Three- and four-blade arrangements present a steady resistance as the tail vane responds to a wind-direction change.

table 2. Power output of spoked-wheel as a function of wind velocity.

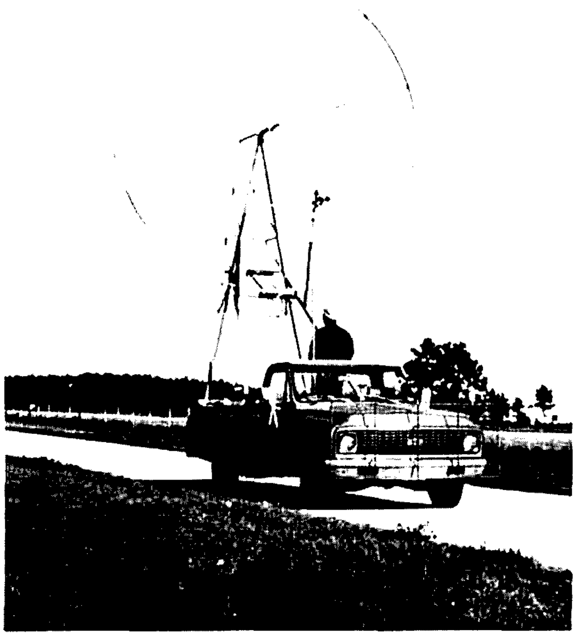
wheel dia., ft.(m)	horsepower, wind speed, mph(km/hr) 10(16)	20(32)	30(48)	kilowatts, wind speed, mph(km/hr) 10(16)	20(32)	30(48)	kW per month, wind speed, mph(km/hr) 10(16)	20(32)	30(48)
7.64(2.3)	0.147	1.18	3.98	0.077	0.615	2.08	55	443	1494
15.28(4.7)	0.589	4.71	15.91	0.307	2.460	8.31	221	1772	5979
30.56(9.3)	2.360	18.86	63.64	1.230	9.840	33.22	888	7087	23,919

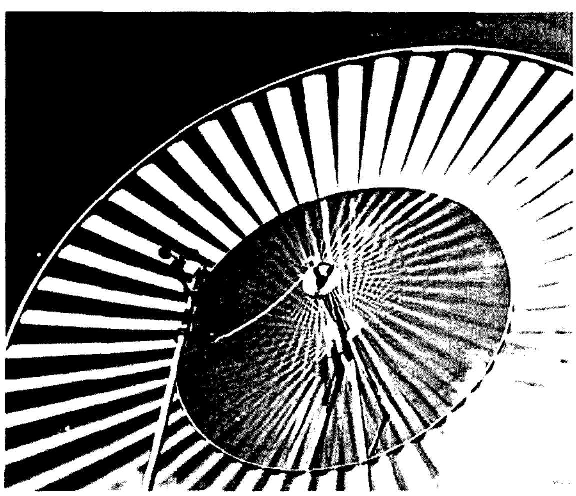
Two lightweight rotators

The Chalk rotator, fig. 4, is unique, effective, lightweight, and starts easily in a light breeze. Early measurements indicate that it can reach an efficiency in the 50 per cent region (recall that theoretical maximum is 59.3 per cent). The Chalk rotator consists of a spoked wire wheel. The structure supports lightweight sheet-aluminum blades shaped in an appropriate airfoil section. The spoked wheel construction provides great strength despite its low weight. For example, a 15-foot (4.6m) diameter wheel weighs about 70 pounds (32kg).

An important advantage of the Chalk construction is that it simplifies gearing to a generator. As an option, it's possible to extract power at the rim. Since the wheel rim speed (comparable with the tip speed of a conventional blade) is high, it may be used to drive a generator directly or may use a very simple gearing system. In fact,

fig. 4. The Chalk rotator.





it's conceivable that the generator field poles themselves can be made part of the rim assembly. **Table 2** shows power output for spoked-wheel wind turbines at various wind speeds. Note that for a small (less than 8 feet, or 2.4m) diameter wheel, 77 watts are generated at a 10 mph (16 km/hr) wind velocity.

An advanced sail wing developed by Princeton University was conceived initially for boat application and eventually as an aircraft wing. Its structure is simple, lightweight, and efficient. Materials are inexpensive and permit a more simplified support structure than conventional blades. A sailwing consists of a rigid leading edge, **fig. 5.** The root section is attached to the rotor hub. Both tip and root are connected by a trailing-edge wire cable, which is fastened to a wraparound sail. The sail is cut in such a manner that its trailing-edge shape is set by the tension of the trailing-edge cable. A taut wing results, with a simple structure. However, the wing deforms and responds to loads in accordance with the wind velocity and angle, developing an effective aerodynamic characteristic. Of importance is its high lift-to-drag ratio. A lift coefficient and gentle stall characteristic compare favorably with the conventional hard wing and blade; it has the same load carrying capability. Furthermore it has the high efficiency of a sophisticated hard blade despite the favorable economics of its structure and support tower. In fact, its weight is such that a two-blade, 25-foot (7.6m) diameter blade is possible before dynamic effects become troublesome. For windmills larger than this diameter, three or more blades are advisable.

A study of wind conditions in the contiguous United States indicates that the maximum ratio between maximum and average wind is approximately 6. Since dynamic pressure increases as the square of the velocity (factor of 36), it's understandable that a windmill must be designed to withstand pressures many times greater that that exerted by the average wind at a given site The effect of strong winds is reduced by braking the windmill or by using a pitch-control system. The fact that the sail blade of the Princeton design is readily deformable results in a twisting component in high wind, which holds the rpm to a safe value.

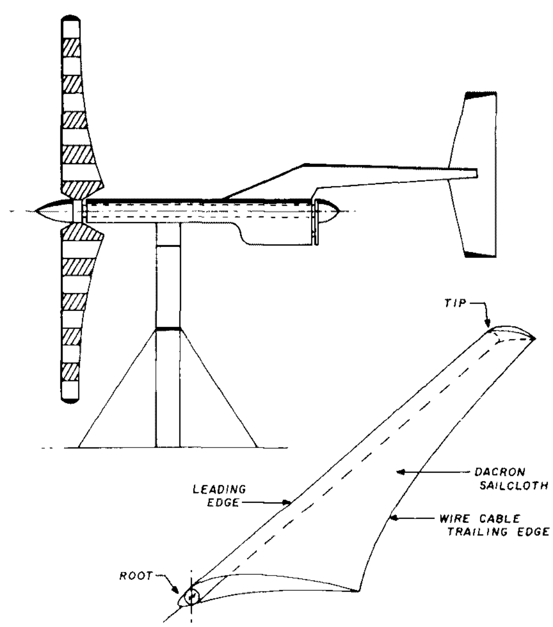


fig. 5. Sailwing generator developed by Princeton University.

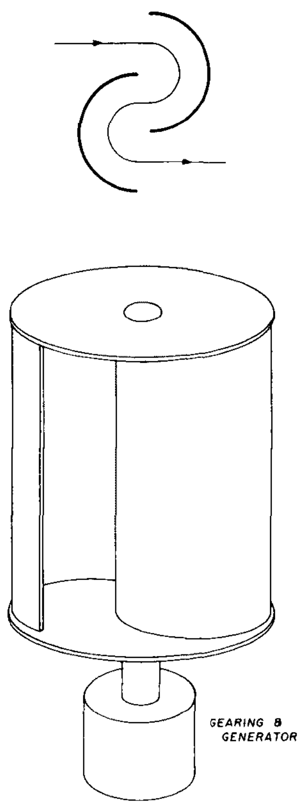


fig. 6. Savonius vertical S-rotor.

The windmills described previously employ structures that rotate about a horizontal axis. Two plans that involve rotation about a vertical axis, although developed many years ago by Savonius and Darrieus (fig. 6 and 7), are being studied and experimented with today. Such rotors respond to wind pressure regardless of wind direction. No vane assembly is needed to orient them into the wind. In general complexity and maintenance are reduced using such a structure. Efficiency is good, and in an area subject to gusting and changing wind direction, output is steadier compared with the horizontal-axis rotor, which encounters loss time during intervals when it's being reoriented by the vane system to accommodate change in wind direction.

The Savonius or S-rotor is a drum-like configuration.* Air striking one of the concave sides of a two-blade arrangement is pressed through the rotor center vent to the back of the convex side, setting up the rotational

*A modified version of the Savonius rotator is sometimes seen on top of buildings where it's used as a ventilator. **Editor.**

pattern. It is a successful wind rotor but much is still to be learned about its characteristics. What is the most efficient and/or effective aspect ratio (ratio of height to diameter)? How does the shape, number of blades, and venting system affect operation?

Every indication shows that the Savonius has a high starting torque, which means that for a general application it will begin to rotate and generate energy at a lower wind speed. Rotation speed is slower but more power can be made available. The slow speed requires a higher gear ratio if the Savonius is to be used as a wind electric generator. However, the mechanical structure can be simplified because the generator can be mounted at the base using a long vertical shaft, **fig. 6**.

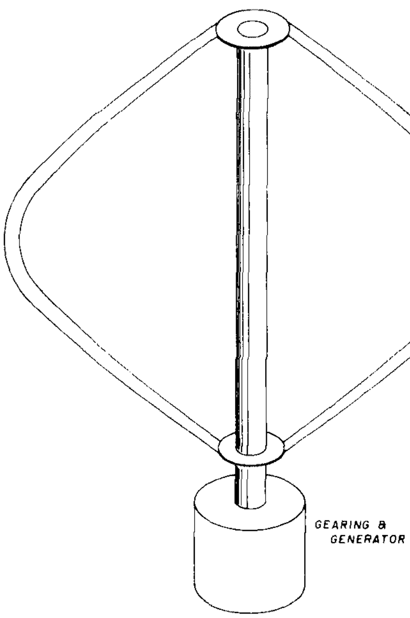


fig. 7. Catenary vertical rotator by Darrieus.

A primitive but very effective S-rotor can be made from discarded oil drums. These are halved and positioned off center. In its simplest form the drum is not reshaped and the venting system remains as is. However, higher starting torque and efficiency may be obtained by making changes to obtain a more ideal vent.

Considerable experimentation is being done with the Darrieus or catenary vertical-axis rotor, **fig. 7**. Tests with both two- and three-blade types are underway. Flexible airfoil blades are used in this construction. Under centrifugal and aerodynamic forces, the blades assume a catenary configuration (bulging at the equator and flattening at the poles). Extensive bracing is not needed and the supporting structure can be quite simple. Bearings at top and bottom and base-mounted gear and generator systems complete the basic structure. A guy-wire system provides additional support for the entire structure. In one experimental unit a 15-foot (4.6m) diameter blade generated 1 kilowatt at a wind speed of 15 mph (24 km/hr). At a wind speed of 30 mph (48 km/hr) the output increased to 8 kilowatts. As an aid during manufacture the blades were made of consecutive straight sections. Fastened together, a reasonable catenary shape can be synthesized. The catenary vertical axis rotor is not self starting. A starting vane or other arrangement starts the initial rotation of the catenary blades.

The vertical-axis wind generator, developed by Sandia Laboratory, combines the Savonius and Darrieus concepts to obtain self starting using a catenary vertical. Two three-bladed Savonius cups (starter buckets) at top and bottom of the power-generating catenary section, **fig. 8**, provide the high torque needed to start the rotating system in a light wind.

wind-generator installation

A 200-watt Winco* wind generator was installed at W3FQJ, **fig. 9**. The specifications of **table 3** indicate that in an area with a yearly average wind speed of 10 mph (16 km/hr) you can expect an average output from the generator of about 20 kilowatt-hours per month. This figure can be somewhat more or less depending on the season. Ten-year average monthly wind speed figures in mph (km/hr) for Philadelphia, Pennsylvania are:

Jan 10.3	(16.6)	July 8.3	(13.4)
Feb 10.8	(17.4)	Aug 7.7	(12.4)
Mar 11.8	(19.0)	Sept 8.0	(12.9)
Apr 11.1	(17.9)	Oct 9.0	(14.5)
May 9.6	(15.5)	Nov 9.2	(14.8)
June 9.0	(14.5)	Dec 9.6	(15.5)

In this area active winds occur in late fall, winter and early spring. Summer wind speeds are lower. Solar panel augmentation is advisable in planning a self-sufficient system. For most amateurs on-the-air activities are limited during the summer months. A 20-kWh-per-month rating at an average wind speed of 10 mph (16 km/hr) provides about 650 watts per day. Not too much power, but enough to run a 100-watt PEP solid-state transceiver almost continuously. Presently amateur radio station self-sufficiency in terms of power is no financial bargain, but it's a pioneering effort and encourages individualism.

As indicated in the specifications, the propeller is a wooden two-blade type, 6 feet (1.8m) in diameter. The propeller hub drives the generator directly; no belts or gear train are used. Generator rpm falls between 270-900 over a wind speed range of 7-23 mph (11-37 km/h).

The site for the wind generator at W3FQJ is on a small rise, reasonably in the clear, at the back of the house. The first step in the installation was the erection of the bottom section of the two-section 15-foot (4.6m) tower. Each leg is supported by a concrete base constructed by pouring concrete into a 2-foot (0.6m) hole dug with a posthole digger. Sixteen 2-foot-long (0.6m) threaded rods, 5/16-inch (8mm) in diameter, support the four base brackets of the tower.

The next step was to attach the wind generator to the top mast section. The top mast section houses a slip-ring

*Winco, Box 3263, Sioux City, Iowa 51102.

assembly that mounts on a platform, fig. 10. Hence the slip-ring and generator assembly rotate as the vane keeps moving the propeller into the wind. Slip-ring and contacts can be seen by opening the slip-ring case, fig 11.

The generator bracket is between two small knobs on the top of the collector-ring cover. This entire assembly was placed atop the lower mast section. Blade and brake

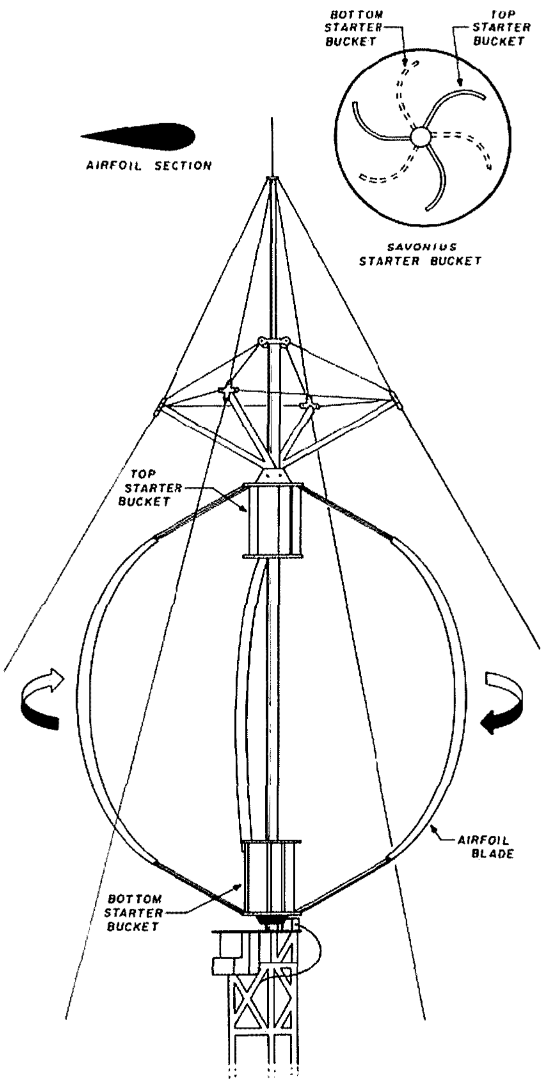


fig. 8. Vertical-axis rotator developed by Sandia Laboratories.

drum were then attached to the generator hub and the vane bolted to the rear generator bracket, fig. 12.

A mechanical governor controls rpm when the wind speed is greater than 28 mph (45 km/h), fig. 12. In normal winds the governor end plates follow a nonresistive circular path as the blades rotate in the wind. At high rpm the centrifugal force opens the end plates, and the resultant resistance holds down blade rotational speed.

A brake rod extends down through tne slip-ring

table 3. Specifications for the 200-watt Winco wind generator.

Tower height	15 ft (4.6m)
Propeller type	2 blade
size	6 ft (1.8m)
material	wood
Gear ratio	direct
Generator	7½" (19cm) diameter
	4 pole
Capacity	200 watts
Approximate maximum amps	14
Approximate maximum volts	15
Generator	
Generator speed range	270/900 rpm
Governor type	22" (56cm) air brake
Propeller speed range	270/900 rpm
Wind speed range	7/23 mph (11/37 km/h)
Average usable kwh per month	
10 mph (16 km/h) average	20
12 mph (19 km/h) average	26
14 mph (22 km/h) average	30
Charge rates	
Revolutions per minute	Amperes
270	0
350	2½
440	6
570	10
700	12
900	14

assembly, and an extension of this rod can be used to brake the generator when desired. Pulling down on the rod causes the brake shoe to engage the brake drum. Braking is recommended when there is a possibility of winds higher than 75 mph (120 km/h). However, to minimize wear, the wind generator can be shut down when it's not being used to charge batteries. In fact, the wind generator should never be operated into an open

fig. 9. Winco generator and Solarex panel installation at W3FQJ.

circuit. A properly placed short circuit at the control box, fig. 13, can eliminate this possibility.

The generator can be used to charge lead-acid batteries, and with suitable circuitry, other types of batteries. One technique is to keep a high-capacity lead-acid battery under full charge and use it, in turn, to charge batteries of lower ratings. The manufacturer of the Winco generator recommends using a 230-ampere-hour battery, which can take good advantage of a sustained high-wind period to accumulate a full charge. If lower-rating batteries are used, care must be taken not to overcharge and you must be ready to change over among batteries.

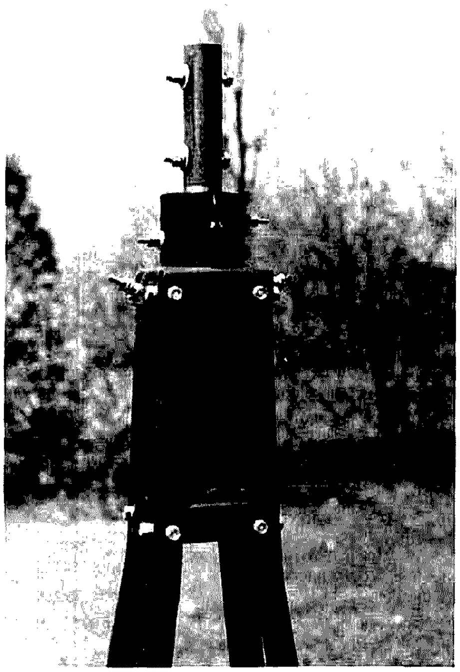


fig. 10. Top of lower section showing slip-ring assembly and platform.

In my installation, the generator is about 500 feet (152m) from the area where the batteries are located and my operating position. I use a small coaster wagon to carry batteries between generator and operating position. From the battery operating location a heavy-duty line runs into my office and electronics bench.

The new 6-volt Gel/Cel 20-ampere-hour batteries are ideal for operating small transceivers such as the Ten-Tec, fig. 14. These batteries are lightweight and can be right in the operating room or can be transported easily when portable operation is desired.

circuit description

The generator- and control-box schematic is shown in fig. 15. The generator is a brush type with a fixed field winding and a rotating armature. When the armature is turning, current flows from brush I through field coils D and C and back to brush H. Current is removed by the

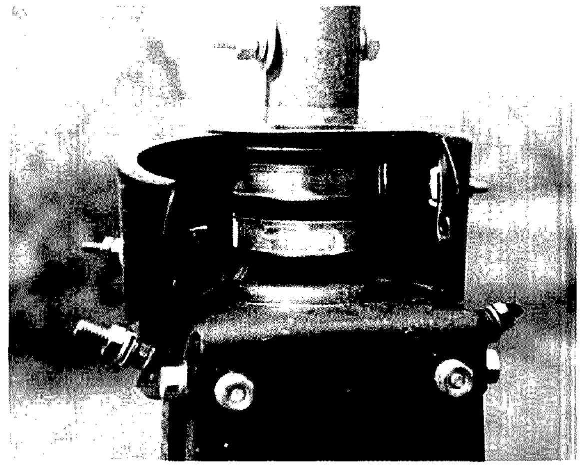
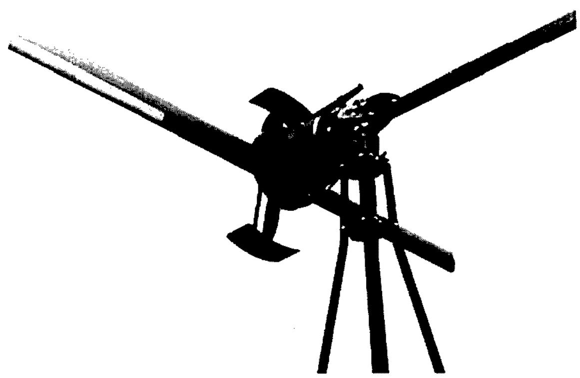


fig. 11. Open case showing slip rings and contacts.

pair of collector rings. Wires connect between the generator and the two terminals of the collector ring cover. These terminals can be seen in fig. 10. Internal connections are made between the collector rings and the two terminals fastened to the top of the tower. From these two terminals, wires are run to the control box. Two no. 10 AWG (2.6mm) wires are used for this connection because, in my case, the control box was attached to the



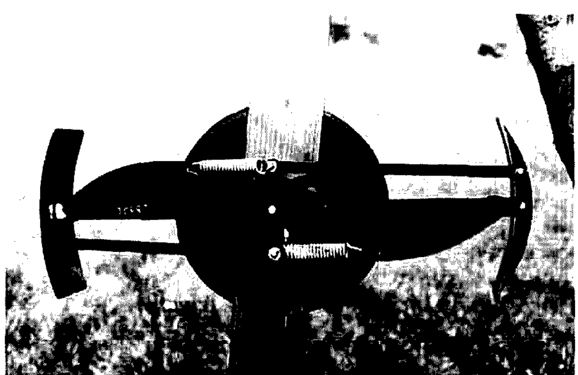


fig. 12. Completed wind-generator (top) and details of propeller-blade, brake-drum, and vane connections (bottom).

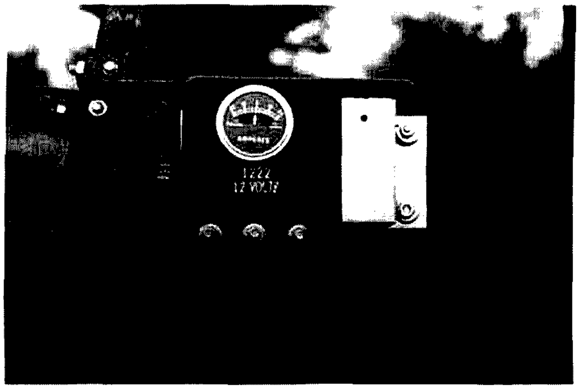


fig. 13. Control box attached to tower. Short circuit (right) is used to prevent system from operating into an open circuit with no load.

mast. If the control box is well separated from the mast, no. 6 AWG (4.1mm) wire is recommended for distances up to 50 feet (15m) and no. 4 AWG (5.2mm) wire for distances between 50-100 feet (15-30.5m). Longer distances require larger-diameter wire. You're working with low voltage and high current; therefore, wire resistance is a very important factor.

The control box houses a charge-discharge ammeter, diode and terminals. The diode prevents the battery from discharging through the generator. However, its polarization is such that charge current from the generator passes with no significant attenuation. The generator is, of course, a dc machine using rectification by brush and commutator. A scheduled-maintenance inspection of the brushes is recommended. Brushes should be replaced when worn.

Note from fig. 15 that the plus side of the generator output connects to the *A-Gen* terminal of the control box. From here the path is through the diode and the ammeter to the control box *+Bat* terminal. The generator negative terminal connects to the control box *F-Gen* terminal. This is the negative side of the battery-charging circuit.

When checking out the wind generator a discharged battery should be connected between the *-Bat* and *+Bat*

fig. 14. Ten-Tec transceiver installation with 6-volt Gel/Cel dry electrolyte batteries.

terminals. The brake can be released and the blade allowed to rotate in the wind. The charging current will be indicated on the meter. In no wind, the circuit can be checked by "monitoring" the generator. In this operation the battery is connected and a short circuit is connected between points B and G (across the diode). This allows the battery current to flow in the generator, which operates as a motor and rotates the blade. In this case the ammeter will read on the discharge side

If the blade is to be rotated without any battery load, a short circuit must be connected between the control box *F-Gen* and *A-Gen* terminals. An open circuit is to be avoided because it results in high voltages appearing in

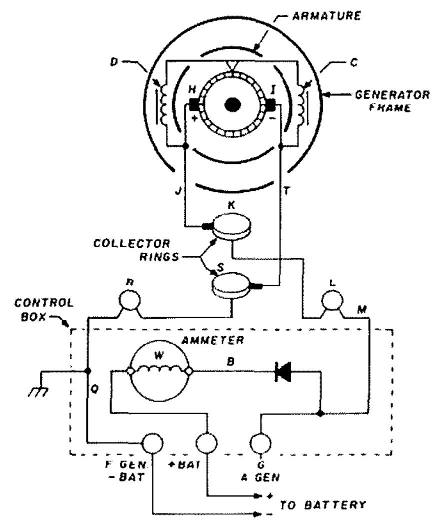


fig. 15. Wind-generator power-supply schematic. Wire size between generator, control box and batteries is important (see text).

the generator and arcing among commutators and brushes. The generator produces a charging current at 7 mph (11 km/h), reaching a maximum in a wind speed of 23 mph (37 km/h). Governing action begins at 28 mph (45 km/h).

Charging current varies with the wind, so it's difficult to keep a charge record unless an ampere-hour meter is inserted in the charging circuit. Therefore a hydrometer is essential in determining when a given battery is fully charged.

As shown in fig. 9 a Solarex* solar panel has been attached to the south side of the tower. This is a 6-watt unit that provides a ½-ampere trickle-charge current. Thus during periods of daylight no-wind conditions a trickle charge can be maintained. A separate control box is being constructed for this panel.

A "much obliged" is extended to Richy Atkinson WA3KHM, Bob Bucher WA3KMW, (now deceased), Heinz Frey, WA3DNZ, Harry Mullen, WA3RLI and Dick Wagner, who aided in the wind generator project.

ham radio

*Solarex, Corporation, 1335 Piccard Drive, Rockville, Maryland 20850.

how to add an inverted V or delta loop to your tower

An easy way to obtain low-angle coverage for 40, 80 or 160 meters using a simple mast extension on your high-frequency beam antenna installation

Over the past few years, and especially since the introduction of the 5BDXCC and 5BWAS awards, numerous articles on antenna systems for 80 and 40 meters have appeared in the amateur magazines. These systems were basically trying to accomplish one purpose: a lower radiation angle on these bands. All too often, however, impractical heights or a considerable amount of real estate were involved.

Presented here is an inexpensive means of extending the height of your tower so you can mount one of the popular inverted-vee or "drooping dipole" antennas for the lower frequencies. The only requirement is a tower of some height with the antenna rotor installed inside the tower.

description

The basic idea is a mast extension at the top of your tower. To this extension a swivel joint is affixed. Above the swivel joint, another 6 to 12 inch (15-30 cm) length of identical mast is mounted, which acts as a mounting base for a low-frequency antenna. A typical setup is shown in **fig. 1**. When erected as shown, everything below the swivel rotates when you rotate your beam. Antennas for 80 and 40 meters, or for that matter, any bands you wish, then serve to guy the installation and keep the short section of mast above the swivel from rotating. Simple? You bet! The cost for the entire assembly, not including antennas, will be less than $20.

design considerations

The upper mast extension is needed so your beam element ends will clear the low-frequency antenna (or antennas) as the beam antenna is rotated. The mast extension length depends on the size of your beam antenna. A typical tribander, such as the Classic 36 or TH6DXX, will require a minimum extension of about 22 feet (6.6m) above the plane of the elements. This is the minimum extension. In practice, about a foot (30cm) should be added to allow for ample clearance. Dimension X in **fig. 2** is the minimum value that will allow the beam antenna to turn freely under a low-frequency wire antenna drooped 45 degrees from the horizontal. Dimension X is determined by simple geometry:

$$X = \sqrt{Y^2 + Z^2} \qquad (1)$$

where X is the clearance height, Y is the distance from the rotator mast midpoint to the end of the boom, and Z is the distance from the boom end to the end of the longest element. Remember that dimension X is the

By Ed Sleight, K4DJC, 4165 Williamsburg Drive, College Park, Georgia 30337

minimum value needed to clear your beam antenna; it's not the maximum value for the mast extension.

construction

The mast extension was made of ordinary *heavy-duty* TV mast, available in 5- and 10-foot (1.5 and 3m) sections. The longer sections are easier to work with and are recommended for the main part of the extension. A 5-foot (1.5m) mast section is best for building the swivel-joint portion. If three 10-foot (3m) sections and one 5-foot (1.5m) section are used, the extension gained will be about 30 feet (9m) above the plane of the beam elements. This will allow about 5 feet (1.5m) of the extension to be enclosed by the rotator mast. While probably feasible, extensions longer than 35 feet (10.6m) have not been tried here.

The swivel joint is the key to the whole system. It may be as simple or as elaborate as you wish. My first one was made with parts from an old tricycle. The latest model was made from a 0.75-inch (19mm) water pipe union joint (Sears part no. 42G 12673) and two short pieces of water pipe screwed into the union joint. The union was tightened snugly, while allowing it to still rotate, then it was secured with a sheetmetal screw to prevent further movement. The water pipe sections, each about 1 foot (30cm) long, were then built up with pieces of aluminum scrap tubing until a force fit was obtained inside the 5-foot (1.5m) mast section.

installation

My low-frequency antennas are fed with baluns. I believe in feeding a balanced antenna with a balanced feed system. The baluns are easily attached to the short section of the swivel and serve nicely as attachment points for the antenna wires. They will also keep the low-frequency antennas separated at the top of the extension. Tape the feedline securely to the sides of the mast extension. This will keep the weight of the feedline more along the centerline axis of the mast and prevent excessive bowing as the mast is raised.

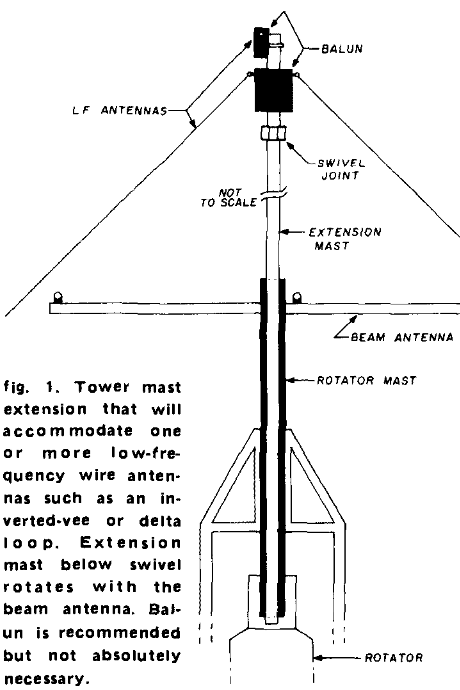


fig. 1. Tower mast extension that will accommodate one or more low-frequency wire antennas such as an inverted-vee or delta loop. Extension mast below swivel rotates with the beam antenna. Balun is recommended but not absolutely necessary.

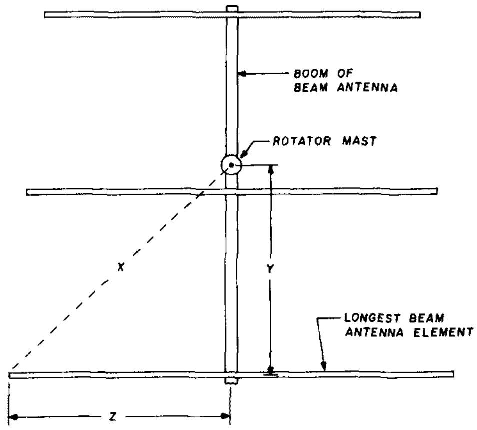


fig. 2. Geometry for determining minimum (see text) clearance of wire antenna that will allow beam antenna to turn freely.

The assembly is easiest to erect if all masting is placed inside the tower, then joined in proper order and fed out the top of the rotator masting. This will usually require removing the rotator. Since the swivel joint will most likely not pass through the rotator mast, it should be fitted in place (with all antennas attached) as the next lower section of masting is pushed upward. Continue feeding the mast extension out the top. When fully extended, secure the base of the extension. I do this by just slipping the rotator back into place.

Tie off the ends of the antennas to obtain the best vertical positioning of the extension mast. Don't worry if the mast leans or bends over slightly while it's being extended. When fully erected, the wire antennas do double duty as guys.

performance

If your tower is in the 50 to 60 foot (15-18m) range, you'll probably have a couple of inverted-vee antennas tied off below the top. With this mast extension, the high-current portion of your antennas will be about 80 to 90 feet above ground (24-27m). On 40 meters, this will lower your radiation angle from about 35 to around 20 degrees.

If your present system is in the 50 to 60 foot (15-18m) range, most of your 80-meter signal will radiate straight up. The system described here will lower that radiation angle to around 45 degrees.

ham radio

five-frequency receiver for WWV

Design and construction of a frequency-standard receiver using a modified transistor broadcast set and an fet converter

For those amateurs whose receivers lack WWV reception, a separate frequency-standard receiver is a useful addition. The special-purpose receiver described here receives WWV on 2.5, 5, 10, 15, and 20 MHz which covers all WWV transmissions except 25 MHz. The receiver gets all of WWVH's transmissions, since WWVH doesn't transmit on 25 MHz.

A block diagram of the receiver is shown in **fig. 1**. The system consists of a modified transistor broadcast receiver and a converter. The transistor radio operates "straight through" for 2.5-MHz reception and serves as an i-f amplifier for the converter, which tunes the higher WWV frequencies. Two crystal oscillators are required, one at 7.5 and one at 17.5 MHz. Each local-oscillator frequency allows dual-frequency reception of WWV, because 10 MHz and 20 MHz are the respective image frequencies when receiving 5 and 15 MHz.

broadcast receiver

The 2.5-MHz section, consisting of a transistor broadcast radio, requires a slight modification to the antenna tuned circuit to make it a fixed-tuned 2.5-MHz amplifier. The modification is simple, because almost all transistor broadcast sets use high-side mixing; i.e., the local oscillator is 455 kHz higher than the received frequency. This means that when the radio is tuned to 1590 kHz, its local oscillator is at 2045 kHz, and the image frequency is 2.5 MHz. Thus 2.5 MHz is easily received by retuning the broadcast radio's antenna tuned circuits to 2.5 MHz. I didn't want the loopstick to function as a 2.5-MHz antenna when receiving other frequencies, so the loopstick was replaced with a small slug-tuned coil. Modification of the broadcast transistor radio is the first step, so it will be treated first.

The broadcast set I used was sold by Magnavox, uses six pnp germanium transistors, and was apparently made in Japan. Newer sets may use npn silicon transistors, and some older GE sets may use npn germanium transistors; however, the conversion will be similar. A typical six-transistor receiver is shown in **fig. 2**.

Tune the receiver to 1590 kHz, remove power, then replace the loopstick antenna (L7 in **fig. 2**) with a slug-tuned coil (L6 in **fig. 3**). Note that the new coil has both primary and secondary windings, so problems of dc isolation are simplified between the broadcast receiver and the rest of the system.

receiver front end

The WWV receiver front end is shown in **fig 3.** Note the use of field-effect transistors. Dual-gate mosfets are used in the rf and mixer stages, while a jfet is used for the crystal-controlled local oscillator. Using dual-gate mosfets as rf and mixer stages simplifies matters, because in each case one gate can be used as a signal-input port and the other for gain control or local oscillator input.

The jfet makes a simpler crystal oscillator than a mosfet because it has a built-in gate-to-source diode, which rectifies oscillator rf voltage and acts much the same as the grid-cathode diode in a vacuum-tube oscillator in establishing grid-leak bias.

The band switch is a five-pole, five-position wafer switch. In the 2.5-MHz position, S1 disables the crystal oscillator and mixer so that only the rf amplifier and the transistor radio are used. Note that L6 is part of a 2.5-MHz parallel-tuned circuit, which is the load for the drain of the mixer fet when the receiver is tuned to 5, 10, 15, or 20 MHz. When the receiver is on 2.5 MHz, L6 becomes the load for the amplifier drain (by means of S1D and S1E). It is for this reason that L1B does not exist.

Rf gain may be adjusted by a 10k pot that varies the voltage on gate 2 of Q1 from zero to +3 volts through a

By **Hank Olson, W6GXN,** Stanford Research Institute, Menlo Park, California 94025

100k resistor. If agc is desired (not included here because it's not known what transistor radio the reader may start with), it would be fed into gate 2 of Q1 through the 100k resistor. The agc would vary this voltage from zero (minimum gain) to about +3 volts (maximum gain), with variations in signal level.

power supply

A regulated power supply for the WWV receiver is shown in fig 4. A three-terminal IC regulator provides +12 volts for the front-end section, and a simple emitter-follower regulator provides +9 volts for the converted broadcast receiver. Some transistor broadcast radios operate on 4 volts (a mercury battery) or even on 6 volts (four penlight cells); it's possible to accommodate these by using a different voltage-breakdown zener in place of CR3. The zener voltage should be close to 0.6 volt more than the desired output voltage of the emitter-follower regulator. A 4.7-volt (HEP Z0405) zener for 4-volt output and a 6.8-volt (HEP Z0409) zener for 6-volt output would be appropriate.

construction

The receiver was built in an LMB W1A box. Looking at the front panel, the power transformer and modified broadcast radio are on the top side of the subchassis at the left. The transformer is the only part of the power supply on top of the subchassis; the rectifier and regulator are on the underside of this subchassis. The broadcast radio (actually its PC board without plastic case) is mounted on the top side of the subchassis. Three or four ¼-inch (6.4mm) fiber or plastic spacers were used in mounting to avoid shorting out PC-board traces. The speaker was remounted on the W1A box front panel (using an appropriate hole size and a piece of perforated metal grille material). Also mounted on this front panel is a volume-control of the same value as that contained in the transistor radio. This control was wired into the radio in place of the original, which was left undisturbed. The original volume control could have been used but wasn't suitable for panel mounting.

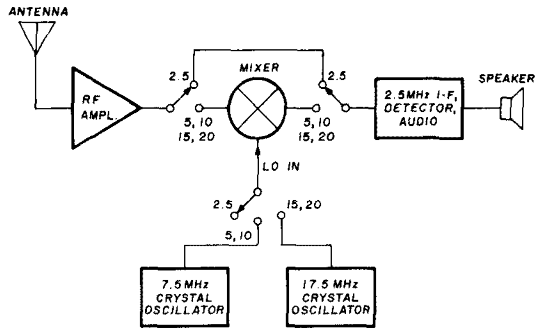


fig. 1. Block diagram of the five-frequency WWV receiver. A transistor broadcast radio operates as a fixed-tuned receiver for 2.5 MHz and as an i-f amplifier for the higher WWV frequencies.

On the right side of the subchassis, a 3¼ x 4¼-inch (82.6x108mm) hole was cut, and a 3-3/4 x 4-7/8-inch (95.3x124mm) piece of dual-sided copper laminate PC material fitted over it. This piece of PC material was secured to the subchassis by six 4-40 (M3) screws. It's far easier to construct rf circuitry on the PC laminate than on the aluminum subchassis, as grounds can be made easily and directly by soldering.

Most front-end circuitry is located on top of this piece of PC board. the main exceptions are the two crystals and coils L2B, L3B, L4B, L5B. These coils should be mounted on the opposite side of the board from L2A, L3A, L4A, L5A to avoid inadvertent coupling and possible oscillation in the rf stages. Because of this simple inductance-isolation method, no further shielding partitions were needed.

The front-end bandswitch was made from a Centralab type PA301 wafer switch with three PA3 sections spaced 1¼-inches (31.8mm). The rf amplifier is located next to the rearmost switch section, the mixer next to the center

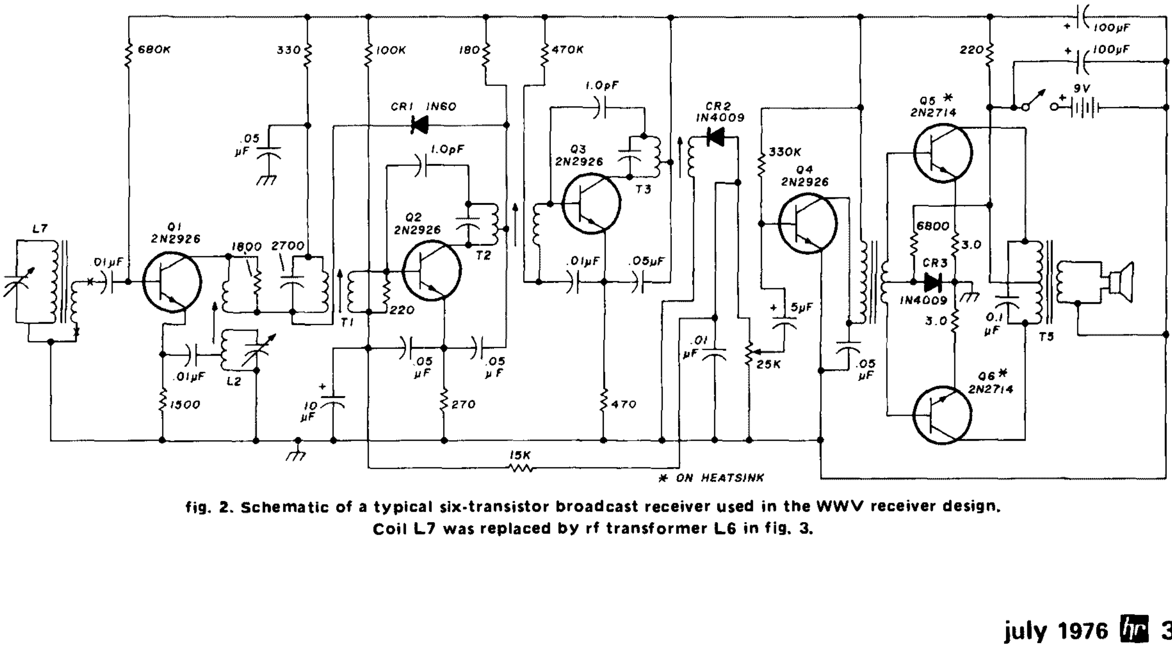
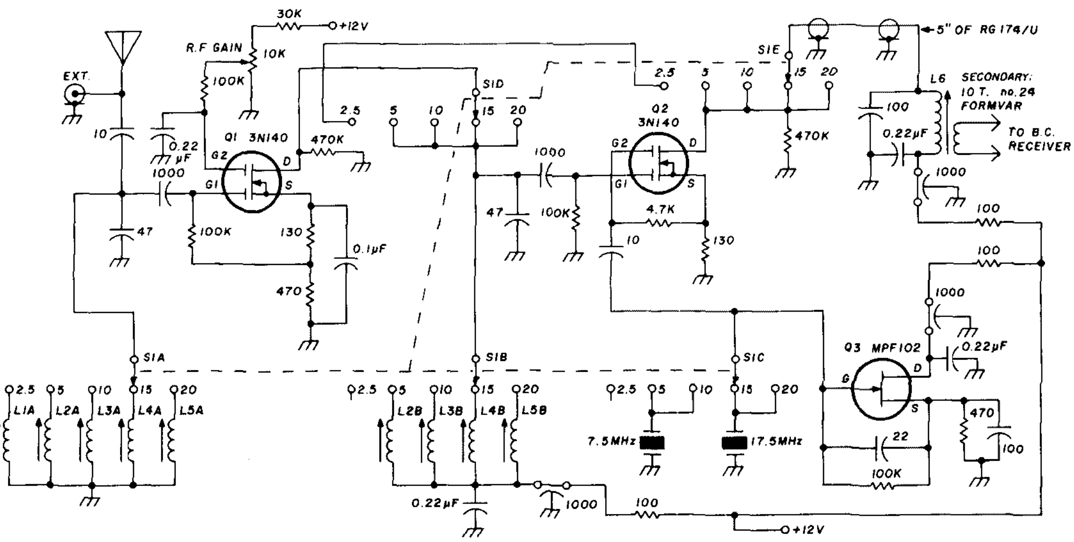


fig. 2. Schematic of a typical six-transistor broadcast receiver used in the WWV receiver design. Coil L7 was replaced by rf transformer L6 in fig. 3.



L1	61 - 122 µH (Cambion X2060-7)	L5	0.83 - 1.6 µH (Delevan 4000-10)
L2	10 - 18 µH (Cambion X2060-4)	L6	28 - 63 µH (Cambion X2060-6), 10 *turn* secondary
L3	2 - 3.7 µH (Cambion X2060-1)	Q1,Q2	3N140, MFE3006, HEP F2004, RCA 40673
L4	1.3 - 2.5 µH (Delevan 4000-12)	Q3	MPF102, HEP802, HEP F0015

fig. 3. Converter for the WWV receiver. Dual-gate mosfets are used for the rf and mixer stages; the local oscillator uses a jfet, which has a built-in gate-to-source diode that rectifies rf voltage.

switch section, and the oscillator next to the front section. Coil L6 is on top of the laminate immediately behind the volume control and to the side of the mixer.

The three +12 volt lines (one each for rf, mixer, and crystal oscillator) emerge through the copper laminate through small 1000 pF feedthrough capacitors, with the 0.22 µF bypass capacitors on the top side. The 100-ohm decoupling resistors are located on the bottom. I strongly recommend that all bypass and coupling capacitors (1000 pF, 0.1 µF, and 0.22 µF) be ceramic disc or Erie *Redcap* types. These have both high capacitance

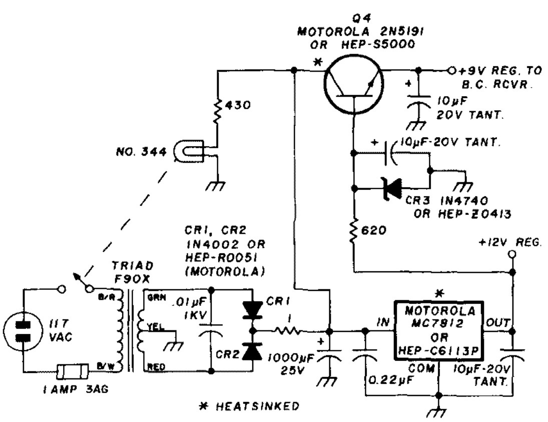


fig. 4. Regulated power supply for the WWV receiver, which provides +12 volts for the converter and +9 volts for the converted broadcast radio. Other transistor radios can be accommodated by using a different zener in place of CR1. The MC7812 *voltage regulator IC is heatsinked directly to the chassis. Transis*tor **Q4 is mounted on chassis with mica washer and silicone grease.**

and low inductance for broadband rf use. The smaller capacitors (10 pF - 100 pF) should be silver-mica.

If the older types of dual-gate mosfets such as 3N140 or MFE3006 are used (which have no gate-protection diodes), the transistors must be handled with the usual precautions. That is, a small no. 32 AWG (0.2mm) bare wire must short all four leads until they are soldered into the circuit. The 470k resistor from each drain of Q1 and Q2 ensures that, during band switching, some dc resistance is always between all elements of each mosfet.

tune up

Tuning the receiver is simple, especially if you have a good signal generator. With the broadcast radio tuned to 1590 kHz, inject a 2.5-MHz signal into the WWV receiver (bandswitch on 2.5 MHz) and maximize the broadcast receiver output near 1590 kHz. Then adjust L1A and L6 for maximum output. Next switch to 5 MHz and peak L2A and L2B; in a similar manner, maximize L3A and L3B on 10 MHz, L4A and L4B on 15 MHz, and L5A and L5B on 20 MHz.

The receiver is designed to use a short piece of wire as an antenna. With a standard 50-ohm generator connected (mismatched) to this antenna, signals of 1 microvolt (1000 Hz, 50% modulated) could be easily detected on all five bands.

references

1. Jim Ashe, W1EZT, "WWV Receiver for $5," *73*, November, 1966, page 20.
2. C. Caringella, "Build a Three-Channel Time Receiver," *Popular Electronics*, December, 1970, page 33.

ham radio

the weekender

shirt pocket transistor tester

The transistor tester described here is useful when troubleshooting circuits. It's also great to take along while shopping the surplus stores for checking unknown devices. Large quantities of older transistors are available at low cost in surplus outlets, many of which are entirely suitable for many projects.

This tester will allow you to grade devices into npn/pnp and good/bad categories. It will also give you a pretty good idea as to whether the transistor is germanium or silicon. The best way to illustrate what the tester can do is to go through a typical test routine. The following procedure is based on the fact that the loaded terminal voltage of the battery in the tester is at least 7.5 volts.

using the tester

Suppose you wish to test a small-signal transistor that has no recognizable markings. Before plugging it into the tester, make sure the function switch is at the *short* position. This test must *always* be made first. Since nothing is known about the device, either npn or pnp can be chosen on the toggle switch. If the meter shows any reading at all, you'd normally reject the transistor without making any further tests. In this case however, if the toggle switch were set to npn and the device under test was a pnp (or vice versa), the meter would indicate a short circuit. (More about this later.)

Move the toggle switch to its other position and check for a meter reading. If the meter still deflects, we no longer care what the device is because it's definitely bad. Do not attempt to make any other tests on a bad device; it's bad for the meter! If the meter reads zero, then we've determined the device type and the fact that it's probably usable. The second step is to determine the quality of the device.

Turn the knob to test for I_{cbo} (collector-to-base current, emitter open). Most modern small-signal transistors exhibit nanoamperes of I_{cbo}. Many older types have microamperes of I_{cbo}. The former will give little or no reading on this meter, while the latter will produce small to moderate readings. We'd normally relegate a transistor with a high I_{cbo} reading to the trash can or give it to some curious youngster, so further testing at this point is really academic. In any case, a transistor that produces a moderate-to-high meter reading is probably a germanium device. Germanium devices typically have I_{cbo} values 10 to 100 times that of silicon devices, which is one reason why most transistors made today are silicon. Since I_{cbo} will increase with temperature and effectively increase the forward bias on the device in a circuit, a condition known as thermal runaway ensues. And if the I_{cbo} reading is moderate to high, be prepared for a high I_{ceo} (collector-to-emitter current, base open) reading on the next test. I_{ceo} always will be greater than I_{cbo} by a factor approximately equal to the current gain (beta) of the device. The modern silicon devices may not show an indication on the meter on this test either.

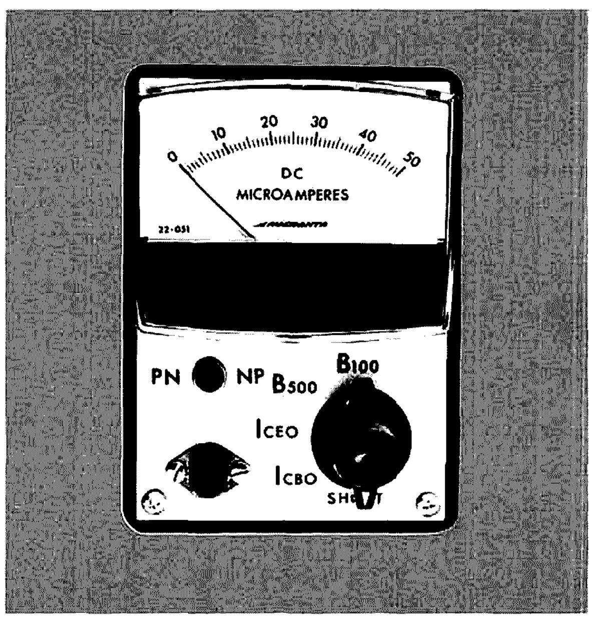
The final test is for beta. The switch selects a high beta range of 500 and a low range of 100. You won't find transistors with a beta of 500 very often, but they do exist. A transistor I like to use is a 2N3391, which typically gives beta readings between 300 and 400. The more usual case for the older types is a beta of 100 or less, and many of the types made for switching applications read quite low.

Having finished this test sequence, we've obtained some pretty useful information about the device under

*A complete parts kit for this transistor tester is being made available in conjunction with this article. For ordering information and prices, write to G.R. Whitehouse & Co., 10 Newbury Drive, Amherst, New Hampshire 03031.

By Dave Cheney, WØMAY, 4808 N. Monroe, Loveland, Colorado 80537

Simple shirt-pocket transistor tester is packaged in small instrument box only 2½" (66mm) wide and 4" (85mm) high.



test. Of course, the most important aspect of these tests is to determine if the device is suitable for our needs. Admittedly the tests don't tell us anything about frequency response, but this requires an entirely different tester. If you really must know whether a device is germanium or silicon, an additional test can be made. Set a volt-ohmmeter or vacuum-tube voltmeter to the one-volt range, and connect the meter test probes between base and emitter with the device plugged into the tester. Select npn or pnp, depending on the device type, and set the function switch to the *B500* position. A germanium device will read 0.2 to 0.3 volt while a silicon will read 0.6 to 0.7 volt.

The short test is really an I_{ces} test, as shown in **fig. 1A**. In this test the full-scale meter reading is 5 mA. Resistor R1 limits the current to about 6 mA, depending on battery condition. This resistor provides some meter protection and limits battery current drain during short tests. A meter reading will be obtained if the device has a collector-to-base or a collector-to-emitter short. To see why an npn device will show a short when the toggle switch is set to pnp (or vice versa), visualize the collector-base junction as a reversed-biased diode in normal operation. If the battery polarity is reversed, the collector-base junction becomes forward biased, which is the case when the toggle switch is in the wrong position.

The I_{cbo} test circuit is shown in **fig. 1B**. The meter reads 50 μA full-scale, as no meter shunt is used. This value of current seems to be a good compromise for checking the collector-to-base leakage of most transistors. The same current range was used for the I_{ceo} test. Some transistors, especially germanium types with high I_{cbo} and beta, will pin the meter. Fortunately, this doesn't happen very often. If you desire a higher current range for this test and still wish to retain the meter calibration, a meter shunt resistor could be connected to set the full-scale value to 500 μA. The I_{ceo} test circuit is shown in **fig. 1C**.

The *B500* and *B100* test positions use the same circuit, except that different meter-shunt values are used. The *B500* test uses the same meter shunt used in the *short* test. On the *B100* test, a meter shunt resistor sets the full-scale current to 1 mA. The basic circuit is shown in **fig. 1D**, but only one shunt resistor is shown.

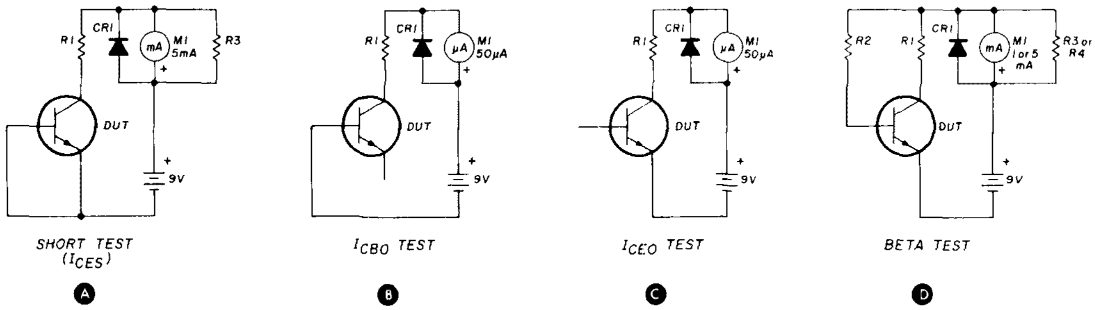


fig. 1. Circuits demonstrating the transistor tester operating principles. R1 is a current-limiting resistor and is in the circuit at all times. R2 sets the base current value for beta tests, and R3 is a meter shunt resistor, which is discussed in the text.

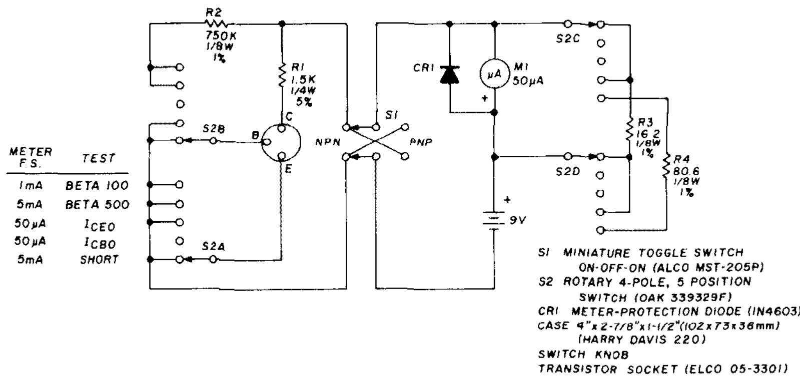


fig. 2. Transistor-tester schematic. Most of the parts are available from Allied Radio Shack stores.

The complete test circuit is shown in **fig. 2**. Note that collector resistor R1 is connected at all times. Its purpose has already been explained. Note, however, that this scheme does not adversely affect the test readings.

The diode across the meter is for over-current protection. Any small-signal germanium unit is suitable. The meter tracking accuracy doesn't compare with that of more costly units but is entirely satisfactory for this use.

The meter that maintains its zero position while in use has good damping. At the present price of about $7.95 it's a good choice for this kind of project. The meter isn't supplied with electrical data on either meter resistance or full-scale terminal voltage. Information such as this seems to be missing from many of Radio Shack's products. I suspect that the low price and lack of information means that the specifications aren't held to close tolerances.

The shunt resistors can't be calculated without having one or the other of the above specifications, so I measured mine. I applied current until the meter read full scale then measured the terminal voltage, which was 80 mV. (I didn't carry the test to its conclusion to see if the full-scale current was truly 50 μA.) Assuming the proper current value, an Ohms law calculation shows that my meter should have an armature resistance of 1600 ohms. If you're so inclined, you can measure your meter terminal voltage and recalculate the shunt resistor values for R3 and R4 using

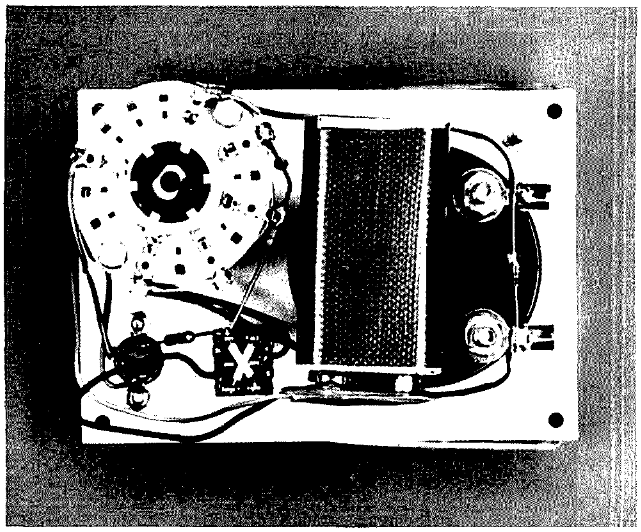
$$R_s = \frac{I_m}{I_s} \times R_m$$

where: R_s = *shunt resistance*
R_m = *meter resistance*
I_m = *meter current*
I_s = *shunt current*

Note that I_s = circuit current minus meter current (I_m).

construction notes

The transistor tester is very compact and the battery is a tight fit. My original intention was to use an 8.4 volt mercury battery (E-146X), which is almost the same size as the common carbon-zinc 9-volt transistor radio battery. The problem is that the two batteries are not exactly the same size. The mercury battery is just slightly thicker and will not allow the front panel to fully close. I mention this because the mercury battery is the most suitable for this application. The mercury battery has a relatively flat discharge characteristic, so that the terminal voltage remains fairly constant until the end of its life. Such a battery will improve the long-term accuracy of the tester. I didn't realize the physical differences between the two batteries until it was too late, and the carbon-zinc cell is used at present. The battery is positioned between the bottom of the case and the back of the meter. The rotary switch in my tester was a junk box item and it's not as suitable as the one in the parts list. The recommended switch uses only two wafers, which provide more room inside.



Transistor tester parts layout. A slightly larger case than used here (see fig. 3) may be used to accommodate a type E-146X 8.4-volt mercury battery, which will improve the long-term accuracy of the tester.

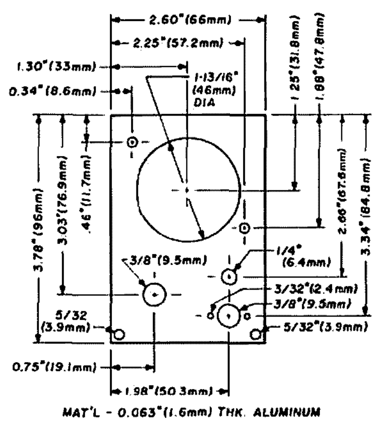


fig. 3. Front-panel layout viewed from the rear. The two 5/32-inch (3.9mm) holes at bottom are located 0.156 inch (3.9mm) in from each side and are for panel-to-case mounting screws.

Those who wish to use my layout may refer to fig. 3 for the front panel dimensions. Note that this is the rear view of the panel, and that the meter-mounting screws are not symmetrically displaced from the center of the meter cutout.

alignment and test

No alignment is needed other than adjusting the meter mechanical zero. The battery should be replaced when its loaded terminal voltage reaches 7.5 V. If the voltage falls below that value, the *B500* test will read low. A simple battery check may be performed as follows:

1. Set the function switch to *short*.
2. Set the toggle switch to either position.
3. Insert a small bare wire in the test socket between the collector and emitter pins.
4. If the meter reads less than 50, replace the battery.

There you have it: a handy-dandy transistor tester you can take to the surplus store in your pocket. If you've read this far, you probably need one too.

ham radio

integrated circuit base-step generator

A straightforward circuit which may be used in building a curve tracer

Several articles have appeared recently describing curve tracer adapters for oscilloscopes. These adapters provide a means of measuring and displaying the current-versus-voltage characteristics of two- and three-terminal semiconductors on an oscilloscope. All of these adapters have two basic sections: a collector sweep circuit, and a base-step generator.

The collector sweep circuit generates a sweep voltage which is applied to the collector of the device under test. This voltage is usually derived from stepped-down line voltage which has been halfwave (60 Hz) or full-wave (120 Hz) rectified. The magnitude of the waveform is controlled by either an adjustable resistive divider or by a variable autotransformer. The circuit also usually contains a resistor for sensing collector current, and/or current-limiting load resistors. The design of this portion of the curve tracer adapter is straightforward and has been adequately discussed in previous articles.[1,2]

base-step generator

To generate a family of curves for three-terminal devices, the adapter must include a base-step generator in addition to the collector sweep circuit. This circuit generates a series of voltage or current steps which are synchronized with the beginning of each collector voltage sweep. These steps are then applied to the base or gate of the three-terminal device under test. The accuracy of this portion of the adaptor is a major factor in the overall accuracy of the curve tracer adapter.

To achieve good accuracy and low parts count, the base-step generator circuit of **fig. 1** was designed using a digital-to-analog, integrated circuit approach.

The resulting circuit has the following features:

1. Calibration of the circuit involves only one adjustment.
2. Accuracy of the circuit is determined only by the accuracy of resistor ratios, not absolute resistor values.
3. A true current source supplies base current.
4. The number of current or voltage steps is selectable.
5. Parts count is low and therefore construction is easy.
6. It is easily adapted to the builder's individual requirements.

circuit

The heart of the circuit is Motorola's MC1406L, a six-bit, digital-to-analog (D/A) converter. Before introduction of this device, most D/A converters were too expensive to be used by amateurs in their projects. Now, a D/A converter is available that is within most amateurs' budgets.* The basic configuration of the MC1406L is shown in **fig. 2.**

A reference current, I_{ref}, is established and flows into pin 12. I_{ref} is given by

$$I_{ref} = \frac{V_{ref}}{R_{12}}$$

The output current, I_o, which flows into pin 4, is an accurate fraction of I_{ref}. This fraction is determined by the digital word present at the inputs of the MC1406L, pins 5 through 10. The output current is given by:

$$I_o = I_{ref}\left(\frac{\overline{A}_1}{2} + \frac{\overline{A}_2}{4} + \frac{\overline{A}_3}{8} + \frac{\overline{A}_4}{16} + \frac{\overline{A}_5}{32} + \frac{\overline{A}_6}{64}\right)$$

For example, if the digital inputs ($A_1 - A_6$) are of the

*Single quantities of MC1406L are available at the price of approximately $5.90.

By Henry Wurzburg, WB4YDZ/7, Motorola Semi-Conductor Products, Inc., Phoenix, Arizona 85008

form 101100, then $\overline{A_1}$ through A_6 would be 010011, and the output current, I_o, would be

$$I_o = I_{ref}\left(\frac{0}{2}+\frac{1}{4}+\frac{0}{8}+\frac{0}{16}+\frac{1}{32}+\frac{1}{64}\right)$$

$$= I_{ref}\left(\frac{19}{64}\right)$$

The output current can range from 0/64 to 63/64 of I_{ref} in increments as small as 1/64, depending on the state of the digital inputs. A set of guidelines is given in Appendix 1 on the proper operating conditions for the MC1406L IC.

If the output of a digital counter is connected to the inputs of the MC1406L, a series of current steps result on its output. This is the technique that is used in the base-step generator of fig. 1. Referring to the circuit, a train of clock pulses is derived from rectified line voltage. My curve tracer adapter used full-wave rectification of the line voltage for the collector sweep. Therefore, the ac voltage from which the clock pulses are derived is likewise full-wave rectified. This voltage is applied to Q1 at point A. The pulses from Q1 are inverted by U1A, and its output is applied to the clock inputs of U2 and U3. These J-K flip flops, along with U1B and U1C, form a synchronous divide-by-eight counter.

The outputs of this counter are then applied to the $A_1 - A_3$ inputs of the D/A converter, U5. I_{ref} is adjusted by potentiometer R1 to be 400 μA and the inputs A_4 through A_6 are tied to +5 volts. With this configuration, the output current into pin 4 is incremented 1/8 I_{ref} or 50 μA, each time the counter is incremented. Note that the minimum increment of current that can be obtained with this configuration is 1/8 I_{ref}, or 50 μA. If current increments other than 50 μA are desired, they can be obtained by appropriate adjustment of R1 and by using inputs A_2-A_4, A_3-A_5, etc. The guidelines for proper MC1406L operation, as set forth in Appendix 1, should always be followed when such a modification is done.

The output of U5 is then fed into a current amplifier composed of U6, Q2, and their associated resistors.

The basic configuration of the current amplifier is shown in fig. 3. The relationship between the input current I_1 and the output current I_2 is given by

$$I_2 = I_1\left(1+\frac{R1}{R2}\right)$$

An important feature of this configuration is that the output current is constant if the input current is constant, independent of the voltage between the output terminals. Note that the accuracy of the current gain is dependent only upon the ratio of R1 and R2, not on their absolute values. This affords the builder the opportunity of obtaining highly accurate current gains without the use of expensive precision resistors.

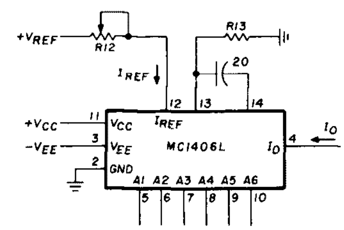


fig. 2. The output current flowing into pin 4 of the MC1406L is an accurate fraction of the reference current flowing into pin 12. The fractional portion of I_{ref} appearing at I_o is determined by the digital word present on inputs A1-A6.

This current amplifier configuration is used in the circuit of fig. 1. Transistor Q2 reduces the amount of current U6 must supply by a factor of Q2's beta. Resistor R9 converts the current steps into voltage steps for fet measurements. This resistor is the only precision part necessary in the circuit. U4 and S1 select the number of steps generated by resetting the counter section at the appropriate point in the count sequence. The circuit power supply is shown in fig. 4. At this point two words of caution are in order:

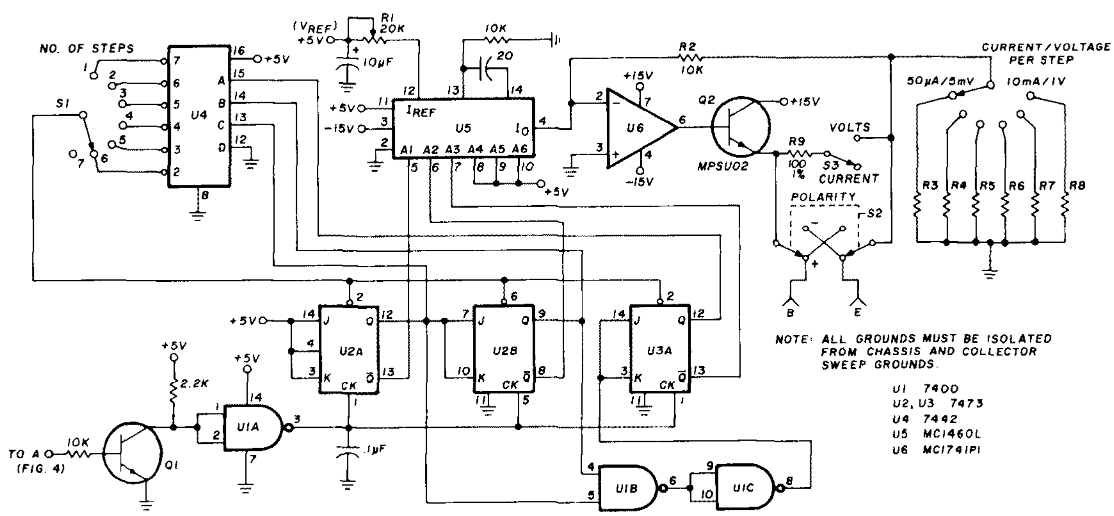


fig. 1. The base-step generator using the MC1406L six-bit, digital-to-analog converter. The values for resistors R2 through R8 are listed in table 1. R1 is a 10-turn 20k potentiometer. Q1 is a general-purpose npn with dc current gain of about 30.

1. Be extremely careful *not* to apply voltage steps to the base of bipolar transistors, or excessive base current will result.

2. The ground for the step generator circuit *must* be isolated from the collector sweep circuit and chassis grounds for proper operation.

construction

Normal construction practices are applicable. However, it is important that the power supply leads be properly bypassed. For the digital portions, a 0.01 μF disc capacitor for every five IC packages is satisfactory. All linear device supply voltages should be bypassed as close to the device as is possible with 0.1 μF disc capacitors.

table 1. Values for R3 through R8 are shown in last column as a ratio of R2 for various gains and voltage- and current-step ratios.

current step	voltage step	gain	R
50 μA	5 mV	1	∞
100 μA	10 mV	2	R2
500 μA	50 mV	10	R2/9
1 mA	100 mV	20	R2/19
5 mA	500 mV	100	R2/99
10 mA	1000 mV	200	R2/199

With the exception of the current amplifier's gain determining resistors, circuit components are noncritical and appropriate substitutions can be made. The accuracy of the circuit is directly dependent upon the selection of resistors R2 through R9.

As was previously mentioned, the gain of the current amplifier is determined by the ratio of two resistors. If an accurate digital ohmmeter or resistance bridge is available, it is a simple matter to "bridge" the resistors necessary to obtain the desired gain. For example, if it is desired to obtain a gain of 100, then R2/R = 99. Since R2 equals approximately 10k, R should be approximately 110 ohms. R2 would be accurately measured and its value noted. Then resistors of 100 ohm nominal value would be measured and a resistor selected whose measured value is closest to being 1/99 of R2's measured value. Highly accurate resistor ratios, and therefore gains, can be obtained in this manner.

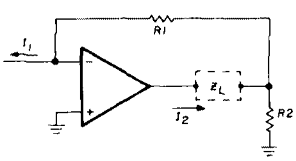


fig. 3. Current amplifier used to buffer the output of the base-step generator follows the configuration shown here. If input current is constant, then output current will also be constant, independent of voltage changes across Z_L.

Of course, this method can not be used if a digital ohmmeter or resistance bridge is not available, or if you don't have a healthy stock of resistors on hand. In this case, precision (1%) resistors must be obtained and used in combinations to obtain the appropriate resistor values.

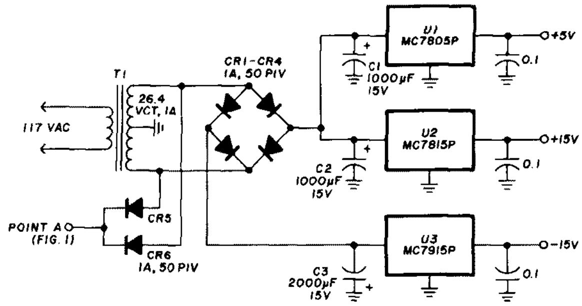


fig. 4. A suitable power supply for the base step generator of fig. 1.

To calibrate the circuit, set switch S1 to step position 7, switch S3 to the VOLTS position, and switch S4 to a midrange position. Connect an oscilloscope to the B and E output terminals and adjust R1 to obtain the correct voltage steps. The circuit is now ready for use.

conclusion

This circuit offers versatility and adaptability. In addition, its accuracy is limited only by the accuracy of the instruments that are used to select the components and to calibrate the circuit. The inclusion of this base-step generator circuit in a curve tracer adapter provides a highly accurate and extremely useful accessory for an oscilloscope.

appendix

The following are some guidelines for proper circuit operation of the MC1406L six-bit, digital-to-analog converter. They are by no means maximum limits, nor are they intended to define all the possible regions of operation. Instead, they are given as an aid for individual design and are appropriate for most circuit configurations. For more detailed information on the MC1406L's capabilities and applications, see the manufacturer's data sheet.

1. Normal operating voltages:

$$V_{CC} = +5 \text{ volts}, V_{EE} = -15 \text{ volts}$$

2. I_{ref} should be equal to 500 μA to 4 mA

3. V_{ref} should be equal or less than +5 volts and well regulated.

4. If V_{ref} is obtained from a logic supply, it should be heavily bypassed close to R12 (fig. 2).

5. R13 should be approximately equal to R12.

6. Voltage on pin 4 (output pin) should never exceed plus or minus 0.4 volt. This may be accomplished by the use of op amp buffering.

references

1. Albert Klappenberger, K3KWX, "An Accurate Solid-State Component Curve Tracer," *CQ*, July, 1974, page 20.
2. Daniel Wright, WA9LCX, "Transistor Curve Tracer," *ham radio*, July, 1973, page 52.

ham radio

readout display

for two-meter digital synthesizers

Dress up your transceiver with a 7-segment display for easy-to-read switch settings

The readout display described here is a great convenience when setting up the channel switches on your two-meter transceiver in the dark or while operating mobile. The display shows the decoded BCD count that controls the synthesizer. If the BCD input to your synthesizer is through front-panel switches, the display makes a nice remote switch-position indicator.

My setup consists of a Heathkit HW-202 and a GLB synthesizer built from a kit. The display is intended for use with the GLB; however, it will work with all the homebrew synthesizers I've seen, of which there are many in this area. I built a three-digit display for $13.00; a two-digit display would be about $8.00. The three-digit display covers all six switches, while a two-digit display would cover the 100- and 10-kHz positions. Since the decoded BCD count that's loaded into the synthesizer is taken from the switches, no mods are required to the synthesizer.

description

The front-panel switches used for frequency selection on the GLB are ten-position rotary switches; i.e., each switch has ten positions with binary-decimal-coded outputs that represent the decimal numbers 0-9. Six identical switches are used; three represent the received frequency (RX), and three represent the transmit frequency (TX). The switches are marked 0 through 9.

The three numbers in a three-digit format represent respectively MHz, 100 kHz, and 10 kHz. Thus if the RX switches are set to 694, for example, and the TX switches are set to 634, you'd be receiving on 146.94 MHz and transmitting on 146.34 MHz. A two-digit format would display the 94 only, for 146.94 MHz. Most homebrew synthesizers have ten-position thumbwheel switches (BCD) that are mechanically different but electrically the same as in the GLB synthesizer.

Since the RX/TX count is loaded into the synthesizer at the same point, the display will normally read out the numbers set up on the RX-selected switches. When you activate the PTT line the readout will automatically display the numbers from the TX-selected switches; this allows three digits to read out all six switches, displaying *only* those in use.

construction

Fig. 1 is the basic circuit and can be used for all three digits. The transistors aren't critical; most any type of npn device will work. The diodes should be silicon switching types. Substitutes for the readouts are MAN-4 or DL-4. *You'll need two identical circuits as in fig.* 1, each of which will be used for the 100-kHz and 10 kHz readouts. The 7404 IC module has six separate inverters. If you build a three-digit display, sections 5 and 6 from the 100- and 10-kHz 7404 modules provide the four inverter for the MHz readout, using pins 10, 11, 12, and 13.

On some synthesizer models the MHz switch is locked from going below 4 and above 7. If you have this type,

By Garry M. Poirier, WB4TZE, P.O. Box 3871, Gastonia, North Carolina 28052

build the MHz display as shown in **fig. 2** using the two unused inverters from the 100-kHz 7404. For the 5 volt supply I used a LM309K regulator (**fig. 3**). Component layout isn't critical. I mounted everything on perfboard except the diodes, which I mounted inside the synthesizer. If you plan to have your display located remotely from the synthesizer, such as dash mounted, small coax should be used for interconnecting wires and you still may have an erratic display while transmitting (this problem is discussed under troubleshooting).

I built my readout into a Minibox, bolted it to the top of the synthesizer, drilled a hole through both, and

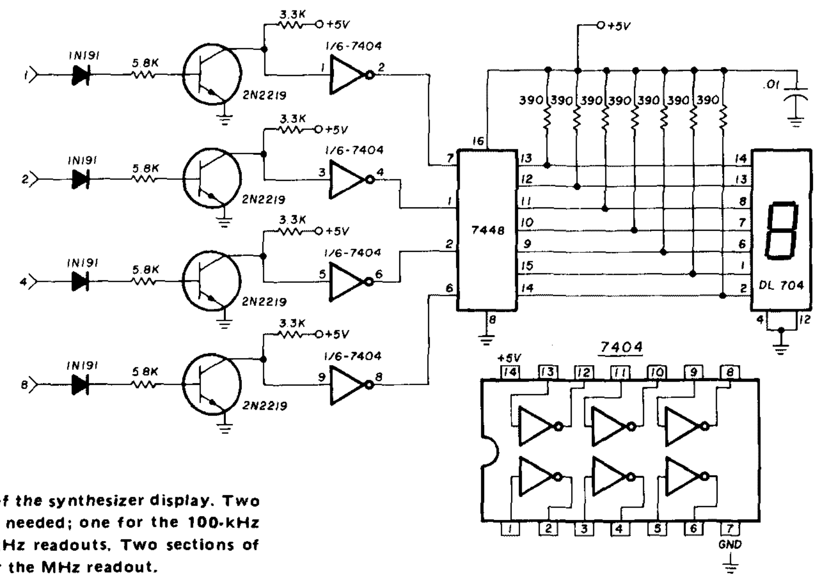


fig. 1. Basic circuit of the synthesizer display. Two identical circuits are needed; one for the 100-kHz and one for the 10-kHz readouts. Two sections of the 7404 are used for the MHz readout.

used standard hookup wire. I cut out the front of the Minibox just enough to view the LEDs, then put a polaroid filter over the cutout.

No modification to your synthesizer should be required to connect the display. However, you must connect the readouts to the proper point for a 1-2-4-8 BCD decoding with automatic switching (see **fig. 4**). Remove the cover from the synthesizer. At the back of the 10-kHz switch you'll see four diodes connected to it, and they in turn connect to four diodes from the lower 10-kHz switch. The point where the diodes connect to each other is the point to make your voltage measurements and to connect the readout diodes.

To find the BCD value of each diode, connect the minus lead of a voltmeter to chassis, turn power on, set both 10-kHz switches to number 1. With the positive lead measure the four diode connections. One and only one connection will be above +3 volts; this point is BCD 1. Move the 10-kHz switches to position 2 and measure again; also do this with the switches in positions 4 and 8. Now that you've determined the 1-2-4-8 diode positions, solder the respective diodes from the display to these

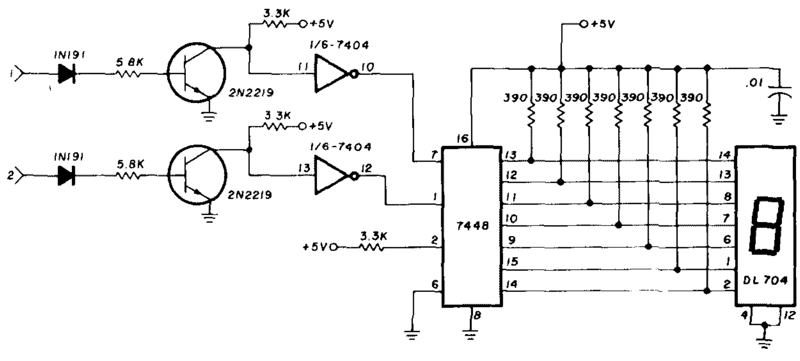


fig. 2. Circuit for limited-range MHz switch. Connect the 2-display diode to the point that measures +4V for both 6 and 7 positions (see text for measurement). The 1-display diode is connected to the other MHz-diode point.

points (watch the polarity). Repeat this operation for the 100 kHz and the MHz readouts. If you have only two diodes on the MHz switch, see **fig. 2**.

test and troubleshooting

When all connections have been made, position the RX switch to use the lower set of switches and turn power on. The display should represent the output *from* the lower switches. Move the RX switch to the upper switches and the display should change to represent the

upper switches. With the RX switch on the upper set of switches and the TX on the lower set, ground the PTT line and the display should change from upper to lower switches.

If a readout is in error from a switch setting and the rig is on the right frequency, then you have made a wiring error, most likely at the diode 1-2-4-8 connections, or you have used a bad component. If the readout

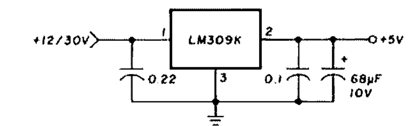
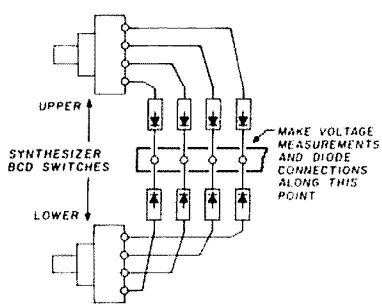


fig. 3. Regulator circuit for obtaining +5 volts. Pin 3 is case ground. The 68-µF capacitor can be any value between 10 to 100 µF at 10 volts.

is correct or erratic and the rig is on the wrong frequency, check for a diode connected backward or a shorted diode. If the reading is correct in RX but in error in TX, you have rf leakage from your rig. Check the reading with the rig terminated into a good dummy load. The diodes isolate the readout from the synthesizer, the transistors bring the signal to a TTL level, the 7404 inverts the signal to the correct polarity, and the 7448 decodes the signal to drive the DL704.

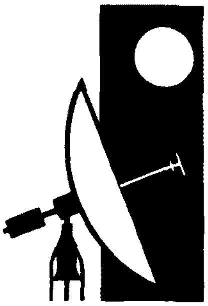
If trouble still persists, such as an erratic readout on transmit, check for rf leakage in and around your set. The shields of all leads from the synthesizer to the transceiver should be grounded at the point of entry into the transceiver. Check for poor antenna connections, high vswr, loose hardware in both rigs, a poor microphone connection, or an oscillating power supply. The readout should be correct and steady, even with a 50-watt amplifier sitting on top of it.

fig. 4. Upper and lower 10-kHz synthesizer BCD switches showing how to connect readout circuit for proper BCD decoding.



The reason for two types of MHz switch hookups is that the GLB has a MHz switch that will not go below 4 (144.00 MHz) or above 7 (147.99 MHz). This prevents out-of-band operation. However, some amateurs wish to use their synthesizer on MARS frequency so they order a band-extended version, which allows the MHz switch to go from zero to 9 (144 to 149 MHz). All homebrew rigs (to my knowledge) will go from 140.00 to 149.99 MHz, or at least the switch is not locked out to prevent it.

ham radio



vhf/uhf techniques

Joe Reisert, W1JAA

matching techniques for vhf/uhf antennas

In my last column, in the May issue[1], I discussed feed systems, feedlines, baluns, matching networks and matching techniques. This month I will concentrate not only on matching techniques but also on easily-built test equipment which can be used to assist in matching and evaluating antenna performance.

More often than not, the station transmitter is used as a signal source to check vhf and uhf antenna performance. This is easy to understand since most stations are equipped with an in-line vswr indicator. However, using the station transmitter for antenna evaluation has several drawbacks. The most obvious problem is that adjustments cannot be readily made with power applied. Another problem which has only recently been mentioned is the possible hazard due to exposure to high-intensity rf fields.[2]

A better technique for measuring and matching vhf and uhf antennas uses a low-power (1 to 100 mW) amplitude-modulated (usually 1-kHz square wave) signal generator. This generator can be either a standard commercial signal generator (such as the Hewlett-Packard 608) or a small solid-state signal source such as that described later in this column. Low power testing eliminates most of the problems associated with using the station transmitter. It is also much less time consuming because adjustments can be made with rf power applied. Furthermore, construction of such an rf generator is simple and straightforward.

Most in-line vswr meters (such as the *Monimatch*) use a microammeter or milliammeter as an indicator. This requires moderate power (more than 1 watt) to drive the meter. Dc amplifiers can be used to increase sensitivity, but they tend to drift and can be quite complex. If a modulated rf source is used, the detected rf signal is ac (rather than dc) so it can be easily amplified to a suitable level to drive a vswr indicator. A typical test set-up is shown in **fig. 1**.

A suitable low-power 144-MHz (10-milliwatt) signal source for antenna measurements is shown in **fig. 2**. It consists of a 72-MHz crystal oscillator followed by a times-2 multiplier. This design follows the general guidelines of reference 3. Experimental tests have shown that *overtone crystal oscillators can be* balky starters. Therefore, I recommend that the oscillator be run continuously while keying the doubler. Simple, off-on (square-wave) keying is preferred. I have also used this technique on a 432-MHz signal source which uses a 108-MHz oscillator followed by two-keyed doublers. Over five years of rough treatment have not caused any problems.

There are several requirements for a suitable rf source. Output power should be constant. This can best be controlled by using a regulated power supply. Battery operation is recommended for field use, but the batteries should be checked periodically for signs of discharge. For best performance in the field, a simple 3-terminal, 12-volt, IC voltage regulator following a 16- to 20-volt battery supply is recommended.

Another rf generator requirement is freedom from load variations (such as the antenna, etc.). This can be satisfied by a 3 to 6 dB attenuator between the generator output and the load. All spurious or harmonic outputs should be at least 30 to 40 dB below the output signal. A double-tuned output filter is usually sufficient (see **fig. 2**). Finally, a shielded box around the generator prevents excessive radiation or signal pickup.

Construction is straightforward. My units are built on a 2- by 4-inch (5x10cm) piece of double-sided epoxy fiberglass PC board which is attached to the inside of the top lid of a Pomona 2901 shielded box. All grounded components are soldered to the copper ground plane. Be sure to remove the paint and protective coating where the box and lid make contact. This insures a well-shielded generator – a must for good measurement.

The 144-MHz rf generator has a CW output of 10 milliwatts and a modulated output of 5 milliwatts. All spurious and harmonic frequencies are 35 to 40 dB

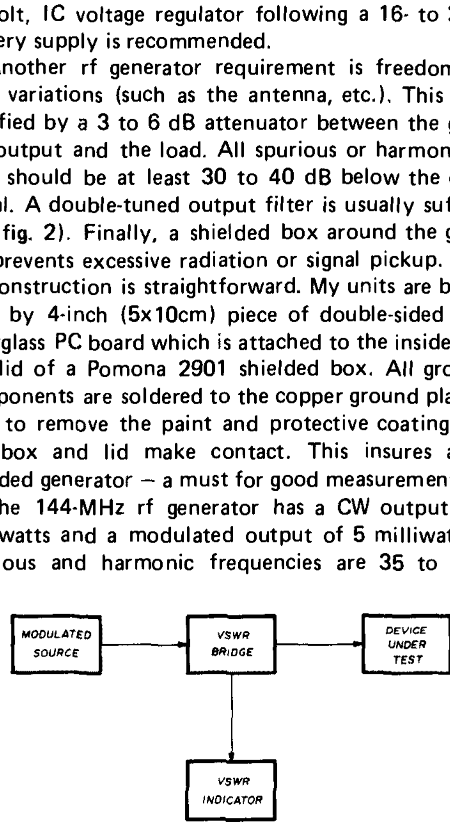


fig. 1. Low-power vswr test set-up. System sensitivity is increased considerably by the use of a modulated rf signal source.

below the output. These characteristics are near optimum for the low-power antenna tests to be described in this article.

You may ask if a separate generator is needed for each band you're interested in. The answer is no. With the exception of 50 and 220 MHz, all major amateur vhf/uhf bands are harmonically related to 144 MHz so it's only necessary to build a 144-MHz generator and

add on amplifiers and multipliers for other bands. Keying the multipliers is not necessary because the 144-MHz source will supply sufficient modulation. Such a scheme is shown in fig. 3. It will be left to the user to provide the circuitry necessary to implement this system.

For several years I have been using a simple multivibrator and series pass transistor as the modulator for my rf generators. At the insistance of K1LOG, I finally updated my circuit to use an NE555 IC. One NE555 timer replaces the multivibrator and its 15 plus components and is easily adjustable from 750 to 1500 Hz. A series pass transistor, while not absolutely necessary, was added since it increased the output by about 2 dB. A complete schematic of the modulator is shown in fig. 4.

This unit is built in a small chassis box with two or more 3- or 4-pin Jones connectors on the output. The output connectors provide ground, +12 volts and keyed 12 volts, thus allowing quick changes or additional units to be plugged in (see fig. 3).

vswr measuring gear

In the May column I mentioned some recommended vswr measuring gear. In this column, I will describe simple homemade equipment that can be used instead of slotted lines, network analyzers and reflectometers. The vswr bridge is a very versatile unit since it can be used to measure various impedances, depending on the reference. Suitable bridges are commercially available from a number of firms including Texscan, Anzac, Telonic and Wiltron. Sometimes these units can be found on the surplus market.

A vswr bridge can be built to work through several GHz, but this requires careful attention to size components, and layout. However, with some care a simple homebrew bridge can be made to work well through 450 MHz. Such a unit is shown in fig. 5. Operation of this bridge is easily understood. If identical loads are placed at J2 and J3, the signals at opposite ends of R3 are equal and in phase, and there will be no output at J4. However, if the impedance of the unit under test at J3 is different from that of the reference load, an output proportional to this difference will be present at J4. The reference load and unit under test can be any convenient impedance value from 25 to 100 ohms. However, the bridge circuit in fig. 4 is designed for optimum performance at 50 ohms.

The values of R1 and R2 are not critical, but both should be the same type and well-matched for best accuracy. This can be easily accomplished by comparing 6 to 10 similar resistors on an ohmmeter and choosing

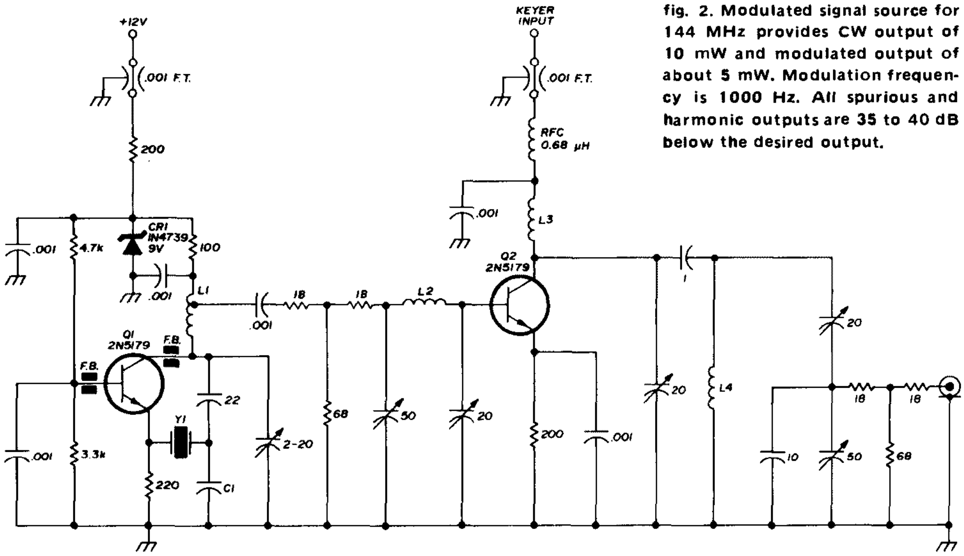


fig. 2. Modulated signal source for 144 MHz provides CW output of 10 mW and modulated output of about 5 mW. Modulation frequency is 1000 Hz. All spurious and harmonic outputs are 35 to 40 dB below the desired output.

C1	as large as possible, consistent with good oscillator starting (100 pF typical)
FB	ferrite bead
L1	9 turns no. 24 (0.5mm) on Amidon T-37-12 toroid core; tapped 3 turns from cold end
L2	15 turns no. 28 (0.3mm) on Amidon T-25-12 toroid core
L3,L4	4 turns no. 24 (0.5mm), ¼" (6.5mm) inside diameter, ¼" (6.5mm) long
Y1	72-MHz, 5th-overtone, series-mode crystal

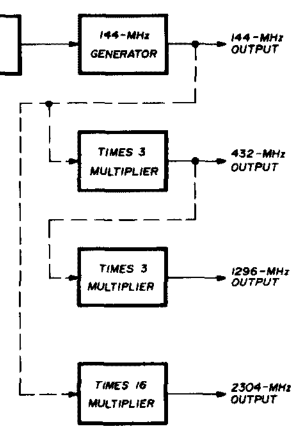


fig. 3. Suggested system for using the modulated 144-MHz signal source as a driver for higher-frequency rf tests. It is not necessary to key the frequency multipliers as the 144-MHz source provides sufficient modulation.

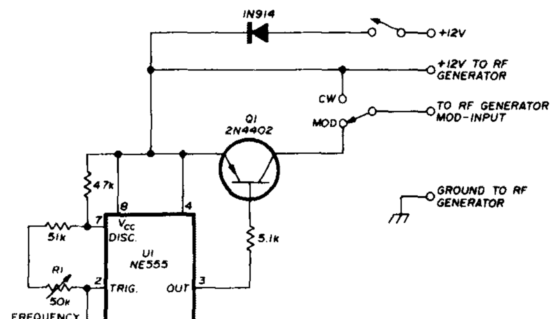
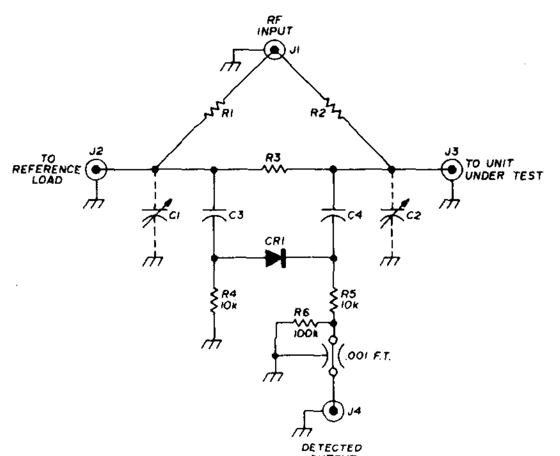


fig. 4. 1000-Hz modulator for the vhf signal source. R1 adjusts frequency from 750 to 1500 Hz. Capacitor C1 should be a Mylar type for stability.

the two which are closest in value. R3 can be one of the rejects since its value is not critical. I would suggest the use of 51-ohm, ¼-watt, carbon-composition resistors. If the bridge is primarily for use at 70 or 75 ohms, R1, R2, and R3 should be changed accordingly for best match to the device under test, but this is a fine point. Capacitors C1 and C2 are small copper tabs that can be added close to J2 and J3 if the ultimate in balance is desired (more about this later).

When building such a bridge, short leads and symmetry are the prime considerations because any long leads or stray capacitance will cause imbalance. A recommended layout is shown in fig. 6. I use a Pomona 2417 shielded box for the enclosure. It is a good choice for the components and type-N connectors.

To test the bridge a modulated rf signal is connected to J1 and an audio detector (described later) is connected to J4. Two identical loads are placed on J2 and J3. The detector output should be extremely low. If not, C1 or C2 can be added to balance out any residual signal. Removing either the load or reference will cause the detected output to rise from 20 to 40 dB or more, indicating proper performance. If two identical loads are swapped from J2 to J3 and vice versa, the detected output should not change.



C1, C2	small capactive tab required for balance (see text)
C3, C4	0.001 μF (small disc ceramic or chip capacitor)
CR1	1N82A or equivalent germanium diode
J1, J4	UG-290A/U BNC connector
J2, J3	UG-58/U type-N connector
R1, R2	47 to 55 ohms, matched (see text)
R3	51 ohms, ¼-watt carbon composition
R4, R5	10k ohms, ¼-watt carbon composition
R6	100k ohms, ¼-watt carbon composition

fig. 5. Homebrew vswr bridge which is suitable for use through 450 MHz. When building the bridge short leads and symmetry are important considerations because of imbalance which can be caused by poor layout. A typical layout is shown in fig. 6.

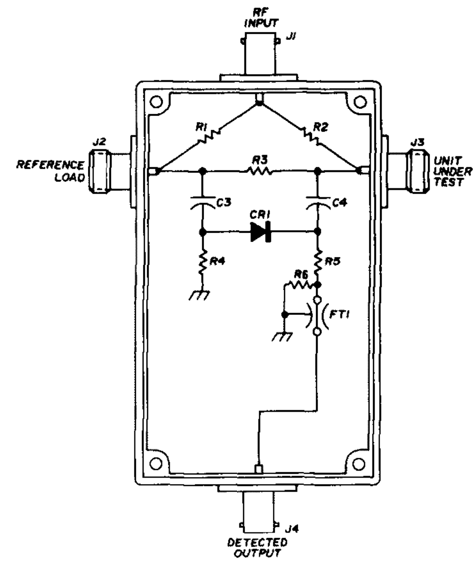


fig. 6. Recommended parts layout for the vswr bridge shown in fig. 5. Enclosure is a Pomona 2417 shielded box. Connectors J2 and J3 are type-N; J1 and J4 are BNC.

Hybrid couplers are also a good choice for vswr measurements and are preferred for operation above 500 MHz where bridges are less accurate. Hybrids are also quite suitable for lower frequencies. Usually quite expensive, recently a low-cost, broad-band, 500 kHz to 400 MHz hybrid coupler became available from Anzac Electronics.* It is mounted in a half-size relay can and is easily adaptable to microstrip circuitry. A schematic and microstrip circuit-board layout for the Anzac hybrid coupler is shown in fig. 7.

If all inputs and outputs of the hybrid coupler are properly terminated, and an rf signal is present at *input 1*, the same signal will be present at *output 1* (less the coupling power) and no signal will be present at *output 2*. The signal level at *input 2* will be below the level at *input 1* by the coupling factor (19.5 dB in this case). However, if the vswr at *output 1* is not 1:1, a signal will

*Anzac model CH137 available from Anzac Electronics, 39 Green Street, Waltham, Massachusetts 02154. *Price is* $13.00 plus tax and shipping.

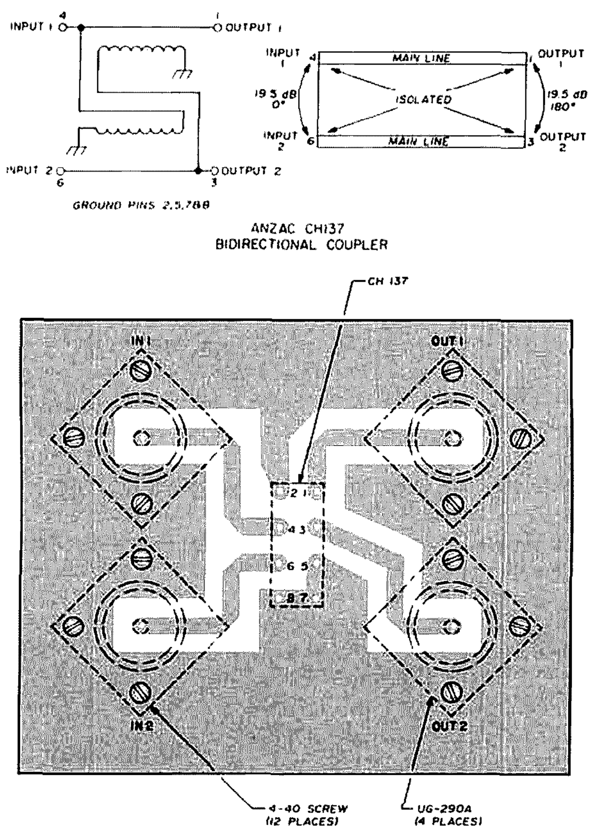


fig. 7. Broadband 19.5 dB directional coupler for use through 500 MHz uses commercially available module (Anzac CH137) and printed-circuit board. The microstrip transmission lines on the 1/16" (1.5mm) double-sided G10 circuit board are 0.1" (2.5mm) wide [clear copper away at least 0.1" (2.5mm) from the microstrip lines]. All connectors are UG-290A/U BNC types; for best performance the shoulder on the rear of the connectors should be removed with a small lathe. Before installing components, on the top side of the board remove copper around pins 1, 3, 4, and 6 of the CH137, as well as around the center conductor of the four connectors.

be present at *output 2* which is proportional to the vswr at *output 1*. The level at *input 2* remains essentially the same.

To perform a vswr test, the signal generator is connected to *input 1; output 1* and *output 2* are terminated in good, nonreactive 50-ohm loads. A 50-ohm detector is connected to *input 2* and the level on the detector indicator is noted for reference. The 50-ohm load at *output 2* is now interchanged with the detector on *input 1* and the detected output should drop considerably (at least 20 dB). Next the load on *output 1* is removed and the device under test is connected. All that is necessary to complete a match is to adjust for minimum detected signal at *output 2*.

This coupler works well through 250 MHz. At 432 MHz there is some imbalance so if it is used at 432 MHz is is necessary to interchange the inputs and outputs, respectively (they are symmetrical, so operation should be identical), to determine which combination gives the best null when properly terminated. In my case *input 2* and *output 2* are better at 144 and 432 MHz.

A narrower-bandwidth stripline coupler with a bandwidth of 10% is shown in fig. 8. I have used this type of coupler through 2304 MHz with excellent results. Operation is identical to that of the Anzac unit. When building the stripline coupler be sure to keep any air gaps between the printed circuit boards to a minimum. Placing 1/8 inch (3mm) thick aluminum plates on top and bottom and bolting them together will keep the air gap to a minimum.

loads

Suitable 50- and 70- to 75-ohm loads are commercially available from a number of sources and can sometimes be found on the surplus market. A suitable homebrew load is shown in fig. 9. I have found that ordinary ¼-watt, carbon-composition resistors to be the least reactive and therefore the most suitable for this application. Half-watt and especially 1-watt units are definitely inferior in this respect. A symmetrical four-resistor load (as shown in fig. 9) has been the best performer – tests have shown this arrangement to work well through 2304 MHz. If a 75-ohm load is desired, the individual resistor values can be changed to 300 ohms.

It is often desirable to have known mismatches to aid in determining the true antenna vswr. A 75-ohm load makes an excellent 1.5:1 vswr reference on a 50 ohm system. Four 100-ohm resistors will make a good 25-ohm load which can be used for 2:1 vswr tests at 50 ohms. In all cases the coaxial connector chosen should be of suitable quality (type-UHF connectors are not recommended above 30 MHz).

Infinite vswr can be tested with a short circuit. Open circuits are not recommended since fringing capacitance will alter the results. A suitable infinite vswr load is also shown in fig. 9.

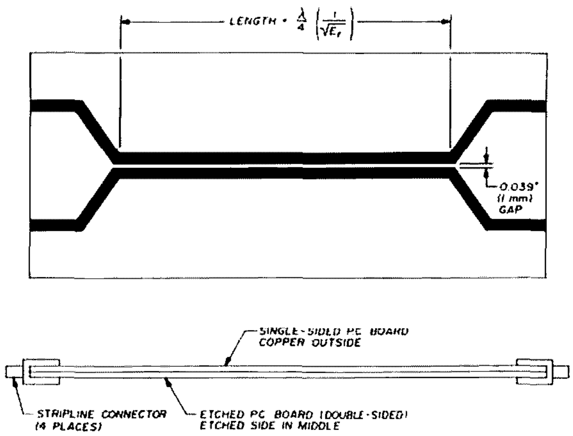


fig. 8. Universal 20-dB stripline directional coupler uses two 1/16" (6.5mm) Teflon-fiberglass circuit boards. Etched board is double copper clad; cover board is single copper clad. The quarter-wavelength of line is shortened by the dielectric constant of the circuit board as shown in the formula (2.5 for Teflon-fiberglass circuit board). Correct length is 4.3" (10.9cm) for 432 MHz, 1.43" (36.5mm) for 1296 MHz, and 1.28" (32.5mm) for 2304 MHz. Bandwidth is about 10%.

detector

A detector is built into the vswr bridge. If a hybrid coupler is used, you must build or purchase your own detector. It should be sensitive and well matched. Point-contact or zero-bias Schottky diodes are preferred because normal Schottky diodes are insensitive at low signal levels unless forward bias is applied. Since the cost of zero-bias Schottky diodes is presently quite high ($25 or more), low-cost point-contact diodes are preferred.

A schematic for a suitable detector using point-contact diodes is shown in fig. 10. The input is a 50-ohm termination and should be similar to the loads in fig. 9. It provides a dc return for the diode as well as providing a load for the hybrid coupler. Typical output is only 50 microvolts at -40 dBm, 5 millivolts at -20 dBm and 30 millivolts at zero dBm. The output is square law (output voltage doubles each time the input doubles) below -20 dBm and linear above -10 dBm. An amplifier is required at low levels.

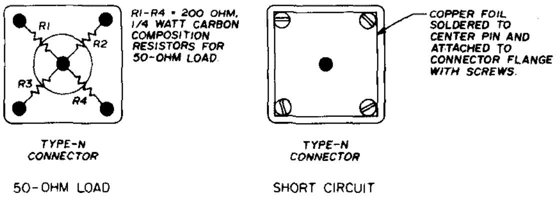


fig. 9. Simple homemade 50-ohm load and short-circuit for use through 2304 MHz. If 75-ohm termination is required, four 300-ohm, ¼-watt, resistors may be substituted.

The detector should have short leads on R1, CR1 and C1. I built mine in a Pomona 2417 box with the input connector on one end, a shield diagonally across the box for the ground return of the feedthrough capacitor, C1, and the output connector on the opposite end. This type of detector is suitable for use to 1000 MHz or so and can be used in many other applications.

vswr indicator

The output of an rf detector is very low at small signal levels. Therefore, an amplifier is needed to drive an indicator such as a meter, and it should be tuned to 1 kHz to work best with modulated signal sources.

Recently many Hewlett-Packard 415 type square law

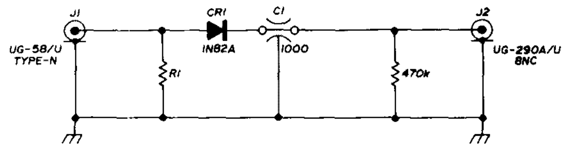


fig. 10. Simple rf detector for 50-ohm systems uses a germanium point-contact diode. Upper frequency range is greater than 2304 MHz. Resistor R1 consists of four parallel-connected, 200-ohm, ¼-watt resistors (see fig. 9).

meters have become available on the surplus market at reasonable prices ($25 to $40). This is an excellent meter to use since it has high gain and is calibrated to match typical detectors. It also has many other uses. A suitable homebrew model has also been described in *ham radio*.[4]

Recently I designed and built a simple, uncalibrated vswr meter that is easily transported and uses a minimum number of components. The circuit, which is shown in fig. 11, consists of a high-gain amplifier, a narrow-bandwidth (100 Hz) selective amplifier tuned to 1000 Hz, and a variable-gain output amplifier which drives a low-cost VU meter. This instrument is ideal for nulling-type vswr measurements, and uses only a single supply voltage. At 9 to 12 volts (not critical) the circuit draws only 5 to 6 mA so an inexpensive 9-volt transistor battery can be used.

matching techniques

Before making any vswr measurements, it pays to set up your vhf or uhf antenna in a clear area, on a tower, or on a wooden ladder pointing toward the sky. The length of the driven element should be set approximately as follows:

$$L = \frac{5500}{f} \quad (inches) \tag{1}$$

$$L = \frac{13970}{f} \quad (cm) \tag{2}$$

where L is the driven element length and f is the frequency in MHz. If the driven element passes through a metal boom it should be lengthened by adding approximately 75% of the boom diameter to compensate for the shortening effect. The length of the driven element is

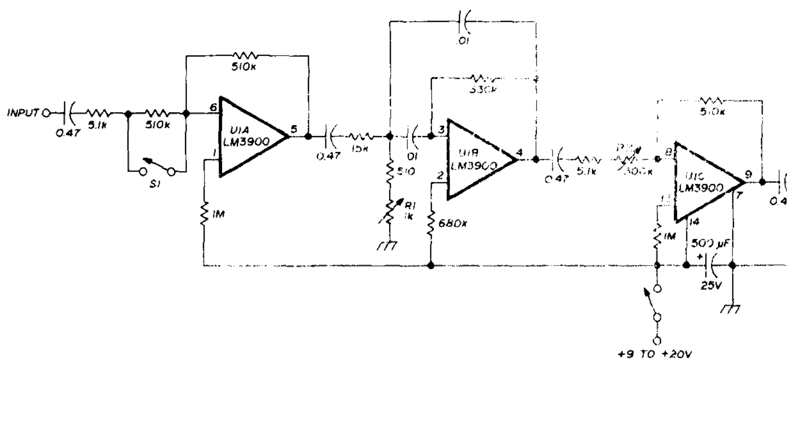


fig. 11. 1000-Hz selective amplifier and vswr indicator. Switch S1, when closed, increases gain approximately 100 times for low-level readings. Potentiometer R1 sets U1B to 1000 Hz, while R2 is used to set the reference on M1, a low-cost VU meter. The 0.01 capacitors should be Mylar types.

not critical for gain purposes and should be tuned for best vswr in conjunction with the matching system.

I have received several inquiries regarding the actual length of a free-space wavelength at vhf and uhf. At the present time authorities are in general agreement that light (and hence radio waves) travels at the rate of 299,792,456 meters per second. Therefore, the correct vhf and uhf formulas for wavelength are as follows:

$$\lambda = \frac{11803}{f} \quad (inches) \tag{3}$$

$$\lambda = \frac{29980}{f} \quad (cm) \tag{4}$$

where f is in MHz. These formulas should clear up any questions on the subject.

table 1. Return loss, dB, vs voltage standing-wave ratio.

return loss	vswr	return loss	vswr
40 dB	1.02	8 dB	2.32
35 dB	1.04	6 dB	3.01
30 dB	1.07	4 dB	4.42
25 dB	1.12	3 dB	5.85
20 dB	1.22	2 dB	8.72
15 dB	1.43	1 dB	17.39
10 dB	1.93	0 dB	∞

Now let's assume we want to match an antenna. If the feed system is similar to the ones mentioned in the last column (delta match, gamma match, etc.), it is only necessary to set up the low-power test setup shown in fig. 1 using an appropriate vswr bridge or hybrid coupler. Initially, a 50-ohm load (or appropriate impedance) can be used to test the matching gear for a null. Then the antenna under test can be substituted and adjusted for minimum vswr. If a complete null cannot be obtained, a mismatch reference can be substituted to determine the actual vswr and to see if any further improvement is required.

If a true square-law detector and indicator are used, the vswr can be determined fairly accurately by measuring the return loss or change between an infinite vswr and the unit under test. Typical values are shown in table 1. If nonlinearities are present (such as would be caused by using the simple indicator in fig. 11), a comparison can be made with known mismatches and the results compared to the test values. It should be noted that coaxial attenuators make excellent mismatches – the return loss is simply two times the attenuation value. An unterminated 3 dB attenuator, for example, yields 6 dB return loss (3.01:1 vswr per table 1).

A 1:1 vswr is a luxury[5] that you may not be able to afford, especially if you operate over a wide band of frequencies (such as 144 to 145 MHz). In addition, aging and weather can often affect the vswr. Therefore, you should strive for 1.5:1 vswr as a goal (approximately 14 dB return loss). This is most easily tested by substituting a 75-ohm load in a 50-ohm system. If the return loss of the antenna under test is better than the 75-ohm load (greater than 14 dB) you can stop making adjustments. If return loss is less than 14 dB, more work is required.

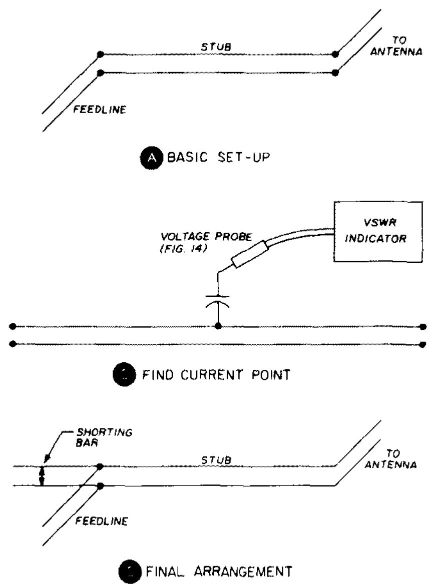


fig. 13. Basic system for matching an antenna to a feedline with a stub. First remove the shorting bar and attach the feedline to the bottom of the stub (A). Feed rf power to the stub and, using a voltage probe, find the voltage minimum (current maximum) which is nearest to the antenna (B). Place the shorting bar at the current maximum point and reconnect the feedline approximately 0.05 wavelength above the shorting bar. Then move the feedline up and down the stub for lowest vswr. When the vswr has been nulled, the null can be enhanced by slightly moving the shorting bar.

Sometimes it is inconvenient to place a balun right at the antenna feedpoint, especially when using stubs. If this happens, you can use tuned lines between the antenna feedpoint and the balun as shown in fig. 12. The transmission line impedance of this extension is not important (250 to 400 ohms is recommended) if the line length is a one-half wavelength or a multiple thereof, using equations 3 or 4. If open-wire line with only a few insulators is used, a physical length of 95 to 98 per cent of the electrical free-space wavelength is recommended.

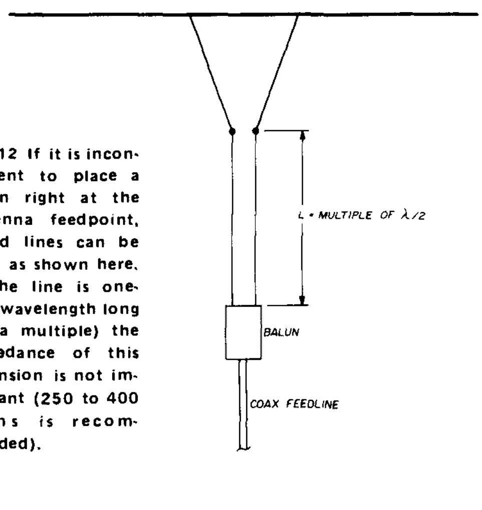
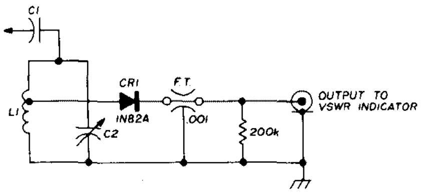


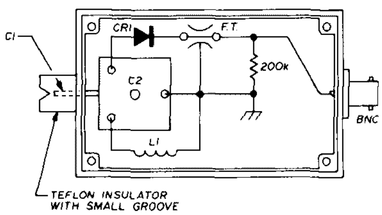
fig. 12 If it is inconvenient to place a balun right at the antenna feedpoint, tuned lines can be used as shown here. If the line is one-half wavelength long (or a multiple) the impedance of this extension is not important (250 to 400 ohms is recommended).

This technique should not be stretched beyond one or two wavelengths because errors are additive, so length becomes more critical.

Although many amateurs are somewhat afraid of matching stubs, they are quite versatile and really *not* that mysterious. Usually a 1.1:1 match can be easily obtained if the right procedures are followed. First, a matching stub should be at least one-half wavelength



CIRCUIT



CONSTRUCTION

C2	1-6 pF for 432 MHz (value depends on frequency)
CR1	1N82A or similar detector diode
L1	2 turns no. 18 (1.0mm), ¼" (6.5mm) diameter, ¼" (6.5mm) long, tapped 3/4 turns from ground end.

fig. 14. Circuit for a sensitive rf voltage probe (A) and method of packaging the circuit in a Pomona 2417 shielded chassis box (B). L1-C2 resonate to the frequency of interest. Capacitance at C1, provided by the gap between the probe and the transmission line should be as small as possible

long. However, if the shorting bar does not fall exactly at the lower end, a match may not be possible. Therefore, a full wavelength is recommended. If this is not possible, an appropriate length of line can be placed between the stub and the device to be matched to place the shorting bar at a convenient place on the stub. This is a handy technique for tight spots (such as the middle of an EME array).

Probably one of the biggest reasons amateurs avoid stubs is the cut-and-try approach that is required to find the proper lengths. However, matching time can be greatly reduced if a few simple tests are performed (such as those suggested to me by W6VSV from the late W6GD). The first step is to determine approximately where the shorting bar is to be placed. To find this point, proceed as follows:

1. Remove the shorting bar.
2. Attach the feedline to the bottom of the stub (see fig. 13).
3. Feed rf power to the stub from a modulated signal source.
4. Using a voltage probe, find the minimum point (current maximum) which is nearest to the antenna. If this point is too close to the antenna end, either shorten the feedline to the antenna or move further down the line to the next current maximum point.
5. Place the shorting bar at the current maximum point.

Next, reconnect the feedline approximately 0.05 wavelength above the shorting bar and move the feedline up or down the stub for the lowest vswr. Then move the shorting bar slightly to enhance the vswr null. Repeat these steps until a suitable match is obtained. The entire procedure is quite simple and takes longer to explain than it does to accomplish!

A simple voltage probe is shown in fig 14. It can be easily built into a Pomona 2417 or equivalent shielded box. In this circuit L1-C2 are tuned to the frequency of operation. When used with the low-power test set-up and sensitive vswr indicator, this provides a very sensitive voltage sensor. When using the device make sure that the hot end of the probe does not touch the feedline. The groove in the Teflon insulator maintains the spacing between the probe and the line and facilitates moving the probe along the line.

gain measurements

By now you have probably guessed that the equipment described here can also be used to measure antenna gain. However, that subject is beyond the scope of this month's column, so will have to wait until another time. In the meantime, if you want to read about antenna gain measurements that can be done with the simple test equipment I have described, I recommend that you read references 6 and 7.

summary

It is hoped that this two-part series on vhf/uhf antenna matching techniques will tempt you to do more work on your antennas. The test equipment described here is a *must* for serious-minded vhf and uhf operators – it will more than pay for itself after a few antenna-matching sessions. And, if you use these techniques, you will no longer have to worry about rf burns or the hazards of rf radiation.

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3. J.H. Reisert, W1JAA, "What's Wrong with Our VHF/UHF Receivers," *ham radio*, March. 1976, page 44.
4. R.M. Melvin, W6VSV, "The SWR Meter," *ham radio*, October, 1970, page 6.
5. M.W. Maxwell, W2DU, "Another Look at Reflections," *QST*, April, 1973, page 35.
6. R.H. Turrin, W2IMU, "Antenna Performance Measurements," *QST*, November, 1974, page 35.
7. R.T. Knadle, K2RIW, "UHF Antenna Ratiometry," *QST*, February, 1976, page 22.

ham radio

carrier-operated relay

for repeater linking

An improved COR using ICs with dual-channel options for linking two repeaters

New concepts and new devices have been developed since the original carrier-operated relay article appeared in these pages.[1] The FCC has approved linking of repeaters, so we've included an addition to the COR that will link repeaters or that can be used for other applications as well.

The basic purpose of a COR is to operate a repeater, so the COR must be as simple and reliable as possible. Other uses for the COR are in a guard receiver for repeater input channels and for speaker muting.

The basic COR circuit (fig. 1A) will handle most repeater requirements. The link, or two-channel COR, shown in fig. 1B can be used for linking two repeaters, for a remote base, or for a guard channel for your repeater. Both circuits have a common front end. Circuit boards are available, including a fully adjustable time-out timer and input-sensitivity control.*

circuit description

Referring to fig. 1, transistors Q1, Q2 are connected as a darlington amplifier for negative-going control signals, as found in a vacuum-tube receiver. The high impedance of the darlington amplifier closely matches the impedance of most tube receivers. For positive-going control signals, normally found in transistorized receivers, make the following changes: R1 to 27k, R2 to 100k, and R3 to 47k. Connect R3 to ground, Q1 emitter to ground, and Q2 base to Q1 collector. The circuit will then respond to positive-going control signals.

U1 is a dual Schmitt trigger that provides positive on-off action. When a control signal is received, ST1 will receive a high from Q2 and a low will appear on its output; this action will set the trigger of timer U2. The low from ST1 is passed to the input of ST2, causing a high to appear on its output, which enables the timer. Time-out is controlled by the setting of R8 and the value of C3.

When the timer starts, its output, pin 3, goes high and energizes the relay. If the input is held long enough, the timer will time-out and pin 3 will go low. However, if the input is released before time-out occurs, the timer output is again driven low and the relay will open. Each time the relay is energized, the timer will have been reset through the action of ST1.

applications

Fig. 2A shows how easy it is to connect this circuit into a receiver and transmitter to control a repeater. Fig. 2B illustrates another use for this circuit. Say you'd like to monitor a repeater or simplex channel for any calls, but you don't want to listen to all the yack-yack going on while you're watching your favorite TV show. Set the timer for about five seconds and when a call comes in, the first few words will be at normal volume. Then, in five seconds, the volume will drop to a low level (determined by the setting of the variable resistor). If the call is for you, simply disable the circuit for normal listening level.

If you wish to control two channels, such as a link repeater or a guard receiver for your repeater, merely add the circuitry shown in fig. 1B. In this configuration we will use the same basic carrier-operated relay shown

*Basic COR board: $4.50; basic COR kit with board: $12.00; link COR kit with board: $19.50. Order from Circuit Board Specialists, 3011 Norwich Avenue, Pueblo, Colorado 81008.

By Robert C. Heptig, KØPHF, and Robert D. Shriner, WAØUZO, Box 969, Pueblo, Colorado 81002

in fig. 1A and add a simple search-lock feature constructed from a single SN7400 and a few other components. Note that R4 (fig. 1A) is changed to 100 ohms, 1 watt for this application.

Gates 1 and 2 of the SN7400 are connected as a simple oscillator, and the output is fed through C9, C10 to gates 3 and 4, which are set up as a dual D flip-flop. Transistor Q3 acts as the lock to stop the oscillator. When no signal is applied to the system, ST1 output is

channels of the link repeater A and B (fig. 2C). Channel A will be set up to receive 94 and transmit on 28. Channel B will receive on 88 and transmit on 34.

When a signal out of the 34/94 machine is received by the link, the signal will be retransmitted automatically on 28 into the other repeater and will come out on 88. As soon as the signal drops out, the search feature will start up again, and if an answer comes back from the 28/88 machine, this signal will be retransmitted through

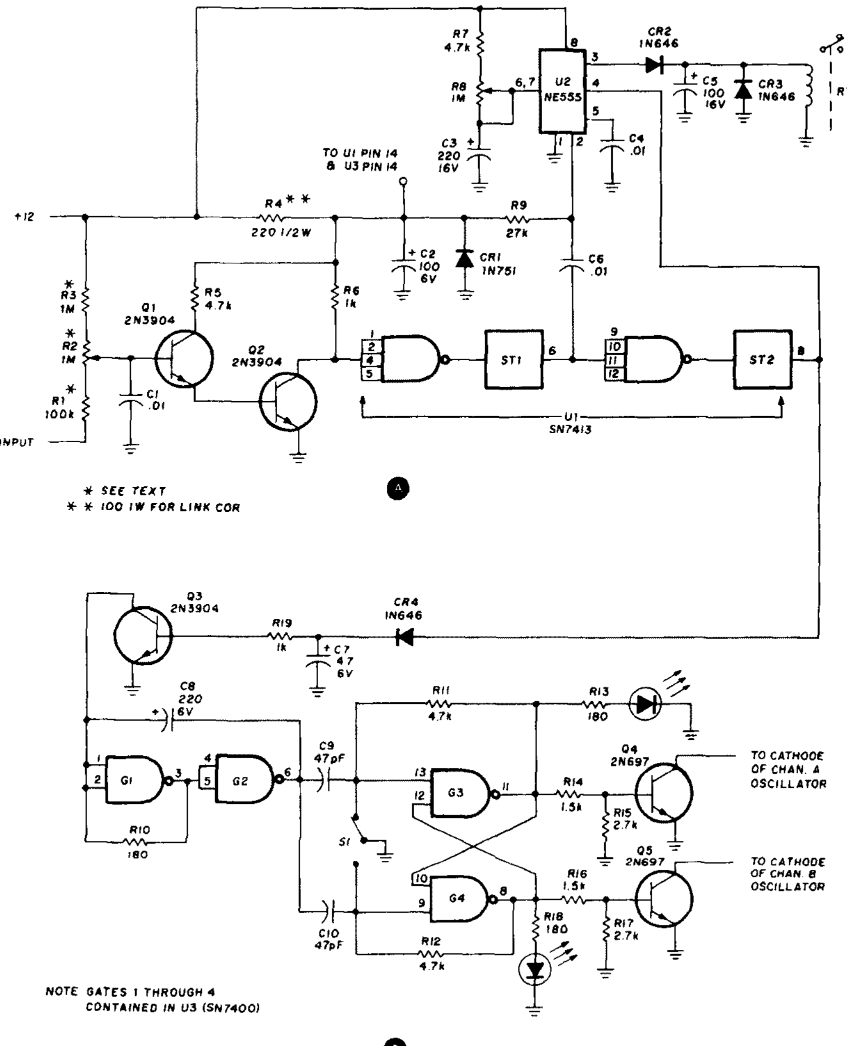


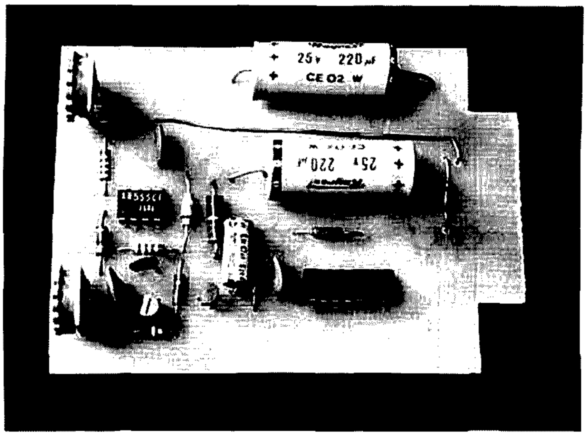
fig. 1. Basic circuit of the COR (A) and simple search-lock feature (B) to control two channels such as a link repeater or guard receiver.

low and Q3 is turned off. This action allows the oscillator to operate and causes transistors Q4, Q5 to conduct and switch on channels A and B respectively.

If a signal is presented to the receiver, the relay will close, causing the repeater transmitter to come up. The oscillator will stop on the channel that was received. As an example, let's say you desire to link a 34/94 repeater into a 28/88 repeater. For simplicity we'll call the two

the 34/94 machine. The action of the basic COR and time-out timer will remain the same as previously described.

How about a guard receiver for your repeater? Refer to fig. 2D. Let's say you have a 34/94 repeater and desire to install a receiver on 94 that won't allow the repeater to come up if the output frequency (94) is in use. Install a second oscillator in your repeater receiver



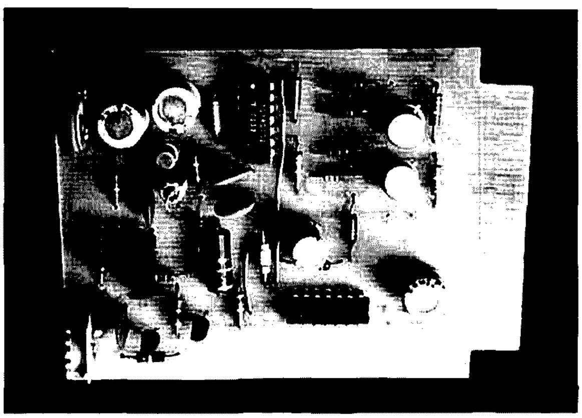
Basic COR component layout. Circuit will handle most repeater requirements. A fully adjustable time-out timer and input-sensitivity adjustment are included.

(channel B) and connect your relay as shown. Diode separation is used to prevent interaction between oscillator and COR. If a signal is not present on 94, the repeater will operate in a normal manner; but if a signal is present then the COR will lock on to channel B and the repeater can't be brought up, simply because the input-channel oscillator (34) is disconnected.

trouble shooting

In the event of trouble use a dc scope or a vtvm to check for the following voltages. (All voltages are given assuming a negative-going signal into the system). First check the supply voltage (+12V) and the voltage to the ICs (+5V). Check Q2 output. It should be close to zero volts with the input open and should rise to near +5 volts with the input grounded and R2 at maximum sensitivity. If not, and Q1 and Q2 are both good, replace Q1, as any slight leakage here will affect the circuit.

To check U1: pins 1, 2, 4, 5 and 8 should be near zero volts with the input open, and pins 6, 9, 10, 11 and 12 should be near +5 volts. These readings should reverse when the input is grounded; if not replace U1.



Added components for a search-lock feature to allow two-channel operation, such as repeater linking.

To check U2: pin 3 should be near zero volts with the input open. Ground the input, and pin 3 should go high (12V); pin 4 will be high (5V). Connect the scope or vtvm to pins 6 and 7. The voltage should rise slowly to about 8 volts, the timer will fire, and pin 3 will go low. If not, replace U2, R7, R8 or C3 in that order.

To check U3: pins 1, 2, 3, 4, 5 and 6 should show a clock pulse of about 2 pulses per second. If not, replace U3, Q3, C8 or R10 in that order until clock is obtained. Probe pins 8 and 11. They should alternately switch from zero to +5 volts. If not replace U3, C9, C10, R11 or R12 in that order until the function is obtained.

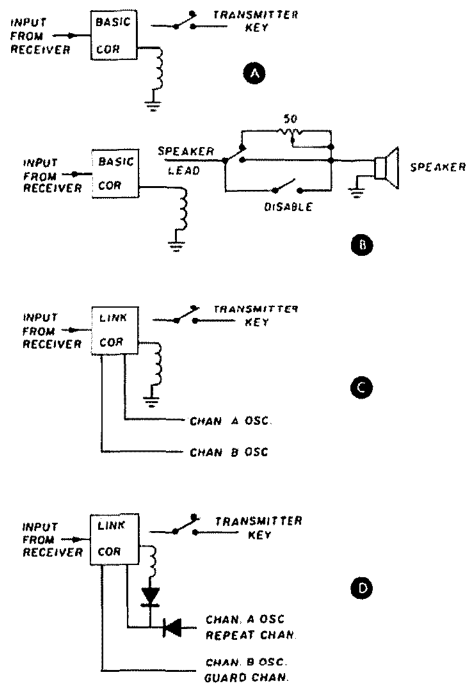


fig. 2. Applications of the COR as a simple repeater (A), for speaker muting (B), as a link repeater (C), and as a guard receiver (D).

By making simple modifications the circuit can be easily tailored to your own requirements. For instance, by increasing the value of C1, a delay activating the system will be noted. Increase the value of C5 to delay the dropout. Decrease the value of C8 to speed up the clock or search rate.

Several of these carrier-operated relays have been constructed and are in use in our area. We've found that others are amazed at the simplicity of the circuit and how well it works for them in their particular application. This circuit is just another example of putting those little plastic centipedes to work for amateur applications.

reference

1. Robert C. Heptig, KØPHF, and Robert D. Shriner, WAØUZO, "Carrier-Operated Relay," *ham radio*, November, 1972, page 58.

ham radio

MICROPROCESSORS

microcomputer interfacing: substitution of software for hardware

A reader who has followed the current literature on microcomputers will frequently encounter phrases such as "hardware/software tradeoffs" or "substitutions of software for hardware." These phrases are strongly indicative of anticipated applications for microcomputers in the near future and do much to explain why industry is so excited about them. In this month's column we discuss how to substitute microcomputer software for hardware.

hardware	Mechanical, magnetic, electronic, electromechanical, and electrical devices from which a system is fabricated.
software	Totality of programs and routines used to extend the capabilities of computers, such as compilers, assemblers, narrators, routines, and subroutines.[1]

In our specific case, software represents the machine-language program stored within the memory of a microcomputer. Hardware represents the specific devices that store, manipulate, receive, or transmit digital information. The microcomputer itself is included in our definition of hardware. The basic point of this month's column can be simply stated:

Through skillful programming, it is possible to substitute machine-level routines and subroutines for specific hardware devices that store, manipulate, transmit, or receive digital information. This activity is called the substitution of software for hardware.

Typical replaced hardware includes knobs, buttons, pulsers, switches, logic switches, clocks, and small memories as well as TTL integrated circuit chips that perform digital functions such as debouncing, sequencing, shifting, adding subtracting, comparing, and logic operations on multibit digital words. Hardware *not* usually replaced includes simple TTL chips such as inverters, flip-flops, gates, latches, three-state buffers, and counters.

Fig. 1 illustrates the basic tools you would employ in the substitution of software for hardware:

1. Programming.

2. The use of synchronized data appearing on the bidirectional 8-bit data bus, D0 through D7.

3. Input and output synchronization pulses called *device-select pulses.*

4. *Interrupts* to the microcomputer.

In an 8080-based microcomputer, you can generate 256 different input and 256 different output synchronizing pulses. If you need more pulses, you can always employ memory I/O techniques as discussed in last month's column. You therefore have an unlimited number of synchronizing pulses with which to coordinate the behavior of almost any type of digital electronic circuit. As you substitute software for hardware, your main tradeoff will be speed of operation. It is useful to remember the following rule:

In the substitution of software for hardware, the key tradeoff is speed of operation. The execution of any computer instruction takes time; the more instructions used, the longer it will take to execute them.

This tradeoff is not as serious as it may seem. Present 8-bit microcomputers are very fast, and future microcomputers will be at least ten times faster. The majority of existing electromechanical machines are slow by digital electronic standards. The human senses cannot participate in activities that require millisecond time resolutions; i.e., in an input/output sense, we are very slow machines.

Table 1 summarizes some of the more commonly

By Peter R. Rony, Jonathan Titus, and David G. Larsen, WB4HYJ

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

encountered situations where hardware such as debounced pulsers, switches, logic switches, and clocks are replaced by simple wire connections, latches, flip-flops, and inverters. We have provided abbreviated versions of the required software. (See reference 2 or previous columns in *ham radio* for details on the generation of the *out n* pulses, where *n* is an octal number between 000_8 and 377_8).

A timing loop is a short subroutine that generates a precise time delay, typically greater than 100 microseconds. As the table shows, the replacement can be made in most cases by the use of one or two different device-select pulses. A pair of *out n* instructions that bracket a timing loop are sufficient, when applied to a SN7474 flip-flop, to produce a monostable pulse of precise time duration. The addition of a second timing loop and a jump instruction, JMP, changes the flip-flop output to that of a variable duty-cycle clock, the duty cycle being controlled by the relative time delays of the two timing loops.

Of particular interest is entry 6 in the table, in which an eight-position mechanical switch or eight individual mechanical switches are replaced by an 8-bit control word strobed into an 8212 chip from the accumulator with the aid of a device-select pulse. This control word is latched by such an action and can subsequently influence the behavior of a rather sophisticated digital circuit. The 8212 chip therefore functions as a *control register* for the circuit. We have directed your attention to this principle because it is now being widely used in an exciting new generation of *interface chips* that reduce the number of wire connections needed between a microcomputer and an external device. The 8255 programmable peripheral interface chip described in last month's column is included in this category.

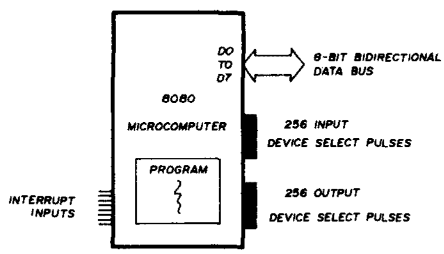
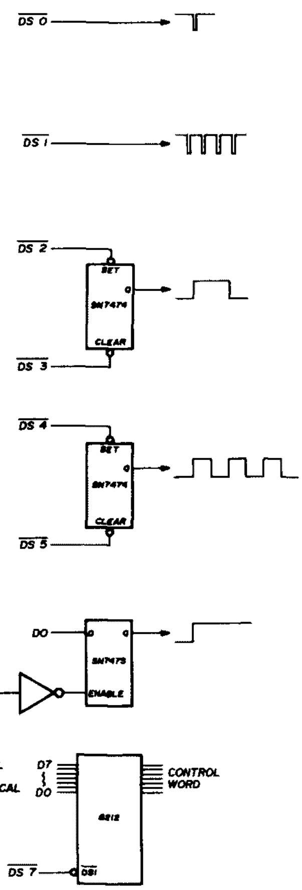


fig. 1. The basic microcomputer tools required for the substitution of hardware by software.

Table 1 provides only a few examples of how hardware can be replaced by simple software with the aid of device-select pulses. Omitted from the table are the more obvious hardware substitutions: arithmetic logic units (SN74181), digital comparators (SN7485), and shift registers (SN74194, SN74198, SN74199). Such chips are

table 1. Some uses for device select pulses in the substitution of hardware by software

SUBSTITUTED HARDWARE FUNCTION	NEW HARDWARE CIRCUIT	SOFTWARE EMPLOYED
1. PULSER, 500 nS MONOSTABLE.	DS 0	OUT 0
2. LOW DUTY-CYCLE CLOCK (500 nS PULSES).	DS 1	OUT 1, TIMING LOOP, JMP
3. DEBOUNCED PULSER, MONOSTABLE, LOGIC SWITCH.	DS 2, SET, Q, SN7474, CLEAR, DS 3	OUT 2, TIMING LOOP, OUT 3
4. VARIABLE DUTY-CYCLE CLOCK.	DS 4, SET, Q, SN7474, CLEAR, DS 5	OUT 4, TIMING LOOP, OUT 5, TIMING LOOP, JMP
5. ONE BIT OF CONTROL INFORMATION FROM A MECHANICAL SWITCH.	D0, D, Q, SN7475, ENABLE, DS 6	MVI A CTRL, OUT 6
6. EIGHT BITS OF CONTROL INFORMATION FROM A SET OF EIGHT MECHANICAL SWITCHES.	D7 … D0, 8212, DS1, DS 7, CONTROL WORD	MVI A CTRL, OUT 7



replaced by microcomputer instructions that add, subtract, compare, and shift the 8-bit contents of the accumulator register.

Although the microcomputer is the most revolutionary electronic device since the invention of the transistor, it is not always obvious how a microcomputer can be used in an amateur radio station. To help lead the way we would like to encourage those that are using microcomputers in amateur stations to drop us a note on how they are being used with the idea of writing a guest column in this section of *ham radio*. Alternatively, you may want to submit a full construction article or even a short note that could be included in one of our regular columns.

references

1. Microdata Corporation, *Microprogramming Handbook*, Santa Ana, California, 1971.
2. *Bugbook III. Microcomputer Interfacing Experiments Using the Mark 80R Microcomputer, an 8080 System*, E&L Instruments, Inc., Derby, Connecticut, 1975 ($14.95 from Ham Radio Books, Greenville, New Hampshire 03048).

ham radio

the ham notebook

micro-TO keyer mods

The schematic of fig. 1 shows a simplified and improved version of this popular keyer. The three transistors that made up the pulse generator were replaced with an NE555V (U1). Normally a free-running oscillator, the NE555V becomes a switchable dot generator by adding CR1, providing an accurate 1:1 ratio.

A variable dash circuit was added using an SN7473 (U2). When U1-12 is high, forming dashes, U2-13 is low, providing a load for U1-7. Voltage then decreases at U1-2, 6, 7; consequently dash length will be extended and may be controlled by the potentiometer shown. In the dot position, CR2 prevents feedback so that dot/space ratio is always 1:1.

The sidetone generator was replaced by another NE555V (U4), which has a tone range of about three octaves. The output easily drives a 3-watt, 4-ohm speaker (no extra speaker is used for the keyer). Still another NE555V (U5) was added to U4 to provide a two-tone oscillator for ssb tuning. A 16-pin IC socket will simplify construction. External components of U5 may be chosen for a suitable fixed frequency, while U4's tone control may be varied to obtain a suitable second frequency and marked for presetting. If you wish to omit the pushbutton switch and U5, connect the 1N914 directly to U4, pins 2 and 6.

The output section consists of Q1 and Q2. The keying bias of my transmitter (an FL-DX500) is -26 volts at 5 mA, easily handled by Q2. Q1 is an inverter for Q2. The resistor across Q1's emitter-collector junction and the capacitor in the base circuit provide a good-sounding on-the-air signal.

The rotary multiswitch connects the receiver output to a speaker or headset. In either position, sidetone can be added by adjusting the volume control while listening to a desired signal. Keying speed and ratio are quickly adjusted, the volume control is turned to zero (no tone), and the transmitter is ready for keying.

The keyer circuit is mounted on a 3-3/16 by 1½ inch (80 by 40mm) perf

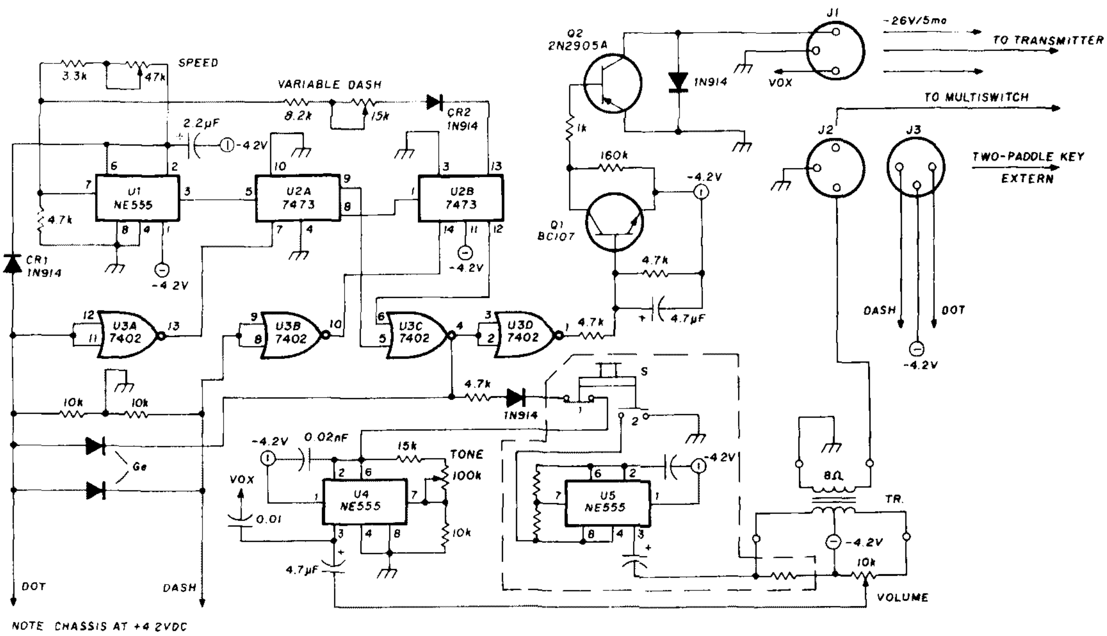


fig. 1. Improved micro-TO keyer. Transformer is a small transistor-radio unit. Keyer speed is about 3-50 wpm; dash ratio can be varied about 5:1. Dot ratio is fixed at 1:1.

board, with space remaining for more components. The transformer is mounted at the inside rear of the box near its jack.

Herbert Seeger, DJ9RP

stabilization of the Ten-Tec KR20 keyer

A nagging problem with the Ten-Tec model KR20 keyer during four years of use has been an intermittent dit when the keyed character was a dah. Persistent checking showed no component failure. Finally, by mon-

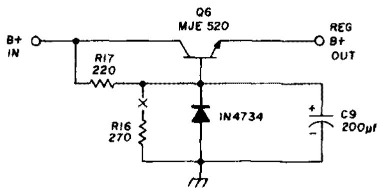


fig. 2. Method for stabilizing the Ten-Tec KR20 keyer.

itoring the output of the power supply, a 0.5-volt shift was discovered in the output voltage while keying. This annoying problem vanished and the shift was eliminated when the 270-ohm resistor between the base and ground of Q6 was replaced with a 5.6-volt zener (fig. 2).

Don Peck, W3CRG

AFSK generator

I recently built the excellent crystal-controlled AFSK generator described in *ham radio** but since I didn't have a +15 volt power supply available, I used the TTL oscillator circuit shown in fig. 3. The crystal is an FT243 that I hand ground to 4589.5 kHz. The 50-pF series capacitor allows the frequency to be trimmed to exactly 4590 kHz.

The output of the TTL oscillator was connected to a divide-by-10 7490 IC and then to pin 1 of U1A in the AFSK circuit.

The output oscillator was connected to a divide-by-10 7490 IC and then to pin 1 of U1A in the AFSK circuit.

I installed two miniature transistor

*Howard Nurse, W6LLO, "Crvstal-Controlled AFSK Generator," *ham radio*, December, 1973, page 14.

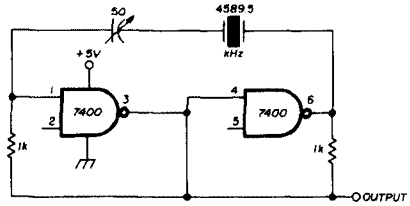


fig. 3. Simple TTL oscillator uses 7400 IC.

transformers at the output of U4B, using the two high-impedance windings connected in series, tuned to 2210 Hz. The output was then connected to the microphone input to my KWM2 through a 22k/100 ohm divider. The output wave-form is very satisfactory and on the air tests were excellent.

Jean Nugues, F8KI

vlf converter

An interesting article appeared in the November, 1974 issue of *ham radio*† describing a very-low-frequency converter with a tuned circuit using magnets and a toroid. The converter shown in fig. 4 uses a lowpass filter instead of the usual tuned circuit so that the only tuning required is with the receiver.

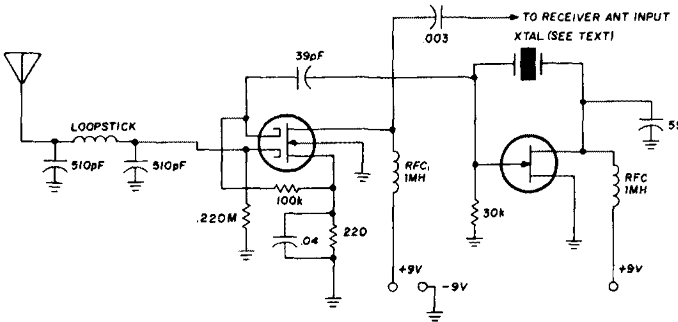


fig. 4. Vlf converter using untuned input.

Despite its simplicity the converter has a measured threshold sensitivity of about 20 microvolts, which is ample for these frequencies. The dual-gate mosfet and fet used in the mixer and oscillator aren't critical. Any crystal having a frequency compatible with the receiver tuning range may be used. For example, I use a 3500 kHz crystal; hence 3500 kHz on the receiver dial corresponds to zero kHz; 3600 to 100 kHz, 3700 to 200 kHz, etc. At 3500 kHz on the receiver all you can hear is the converter oscillator, and vlf signals start to come in about 20 kHz higher.

R. N. Coan, W3CPU

†Guenter Ruehr, OH2KT, "Tuned Very Low-Frequency Converter," *ham radio*, November, 1974, page 49.

tube shields

Many older pieces of gear, such as my 75A4 receiver, use tube shields to isolate various stages. These shields can cause instability if they no longer make good contact with the tube socket.

While this is easily cured by cleaning and by deforming the shield slightly to insure a tight mechanical connection to the socket, a far better course is to replace the shields with one of the more modern tube shield designs.

The old style shields were ineffective at best in helping cool the tubes and often actually caused envelope temperature to rise in local areas, leading to reduced performance and shortened tube life.

The modern shields are easily identified — they are generally black on the outside and feature fluted spring-metal fingers of one sort or another on the inside. These shields substantially lower tube temperature while retaining the isolation characteristics. The old style shields, on the other hand, usually have a metallic finish, and lack any form of heat dissipating fingers on the inside.

Replacement of the old style shields with the more modern variety will give a worthwhile improvement in overall tube life, and will help the tube maintain its new specs longer. The investment is a modest one and well worthwhile.

Bob Locher, W9KNI

Don Roehrs, President of Signal One, and the CX-11 grand prize.

1976 ham radio sweepstakes winners

WA∅QZW is the grand prize winner — eight others win either Atlas transceivers or loom IC-230s

Ham Radio's seventh annual Sweepstakes was by far the biggest ever with nine very happy winners and a very tired staff, both here, and at our local Post Office.

Well over 30,000 entries were received this year, along with many questions about the new Signal/One CX-11, the Icom IC-230 and the Atlas 210X transceivers.

Certainly the greatest interest was focused on the grand prize of the new Signal/One transceiver. Just what is there in that magic box to make it cost $4000? When Don Roehrs, President of Signal/One, personally delivered this prize we were almost as eager as its winner to see what it was all about. We certainly were not disappointed, to say the least.

The CX-11 represents a complete redesign of the earlier models, from a new front-end design in the receiver to a brand new, solid-state final amplifier. Virtually every stage of this intricate radio is either completely new or extensively revised. When you look inside you find almost nothing that looks familiar. The painstaking care that was put into this package should really pay off in both performance and reliability.

We've talked with Randy Powell, WA∅QZW, the winner of the Signal/One after he received it and he could hardly believe what an exciting and complete package he had won.

Next on the winners list were the four lucky recipients of the Icom IC-230 synthesized two-meter transceivers including Glen Galati, WB∅AXK; Dave Mitchell, WA3CPC; Bob McCarthy, WA1UVX; and Helen Haynes, WB∅HOX.

The IC-230 has really made quite a name for itself in the past couple of years. Using a phase-locked-loop synthesizer it covers all the standard 30-kHz repeater pairs and is easily adapted to the 15-kHz "split" channels.

Perhaps the most exciting part of this radio is its very sensitive receiver which includes a five-section helical resonator type front-end filter to insure an absolute minimum of intermod problems. When all of this is put into a package no larger than most crystal-type rigs you end up with one of the most popular 2-meter fm rigs on the market today.

The final group of winners received Atlas 210X transceivers. Included on this list were Chester Koziol, WA2BGS; Robert Trotter, K7VQG; Harry Newport, W∅JDP; and Herb Frosell, K2IB.

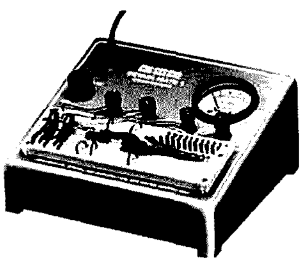
The versatile Atlas features 200 watts dc input to an all solid-state, broadband, no tune-up final. It operates on 80 through 10 meters and offers one of the best receivers available. Extremely small in size and running on 12 Vdc the Atlas is one of the best mobile rigs around today; with an ac supply it is easily adapted to home-station use.

This year's contest was certainly the most work yet (and the most fun yet) for all of us here at *ham radio*, and we certainly want to thank all of you who gave the time and effort to enter. We'll look for *your entry* next year when we draw the grand prize.

ham radio

NEW products

laboratory-grade test instruments



Continental Specialties Corporation has developed three new test instruments to aid the professional engineer, student, or hobbyist in solving electronic design problems. Called Design Mates, these instruments are not kits but are completely wired, fully tested, and ready to use. They are available for immediate delivery from local distributors or from the manufacturer.

Design Mate 1 allows you to build any circuit using no. 22 AWG (0.6mm) solid hookup wire for connecting discrete components, including transistors and ICs in TO-5 or dual in-line packages from 8 - 40 pins or more. No solder is needed for connections because components fit into an appropriate socket and bus strips. Easy in, easy out for circuit testing. Also included is a regulated power supply (5 - 15 volts dc up to 600 mA). You can monitor supply voltage with a built-in voltmeter or use the meter to monitor voltage on the circuit under test. Design Mate 1, wired, tested, and ready to use sells for $49.95.

Design Mate 2 is a full function generator designed for troubleshooting and circuit-design testing. It features three waveforms and a short-proof output amplifier, which provides variable signal amplitudes and constant output impedance. It can be used to test audio amplifiers, operational amplifiers, and prototype circuit designs. It's versatile enough to handle complex industrial electronic design problems. Complete with instruction and operating manual, Design Mate 2 is priced at $64.95.

Third in Continental's new instrumentation series is Design Mate 3, a low-cost R/C bridge. Takes the guesswork out of deciphering component values with unreadable markings. You can measure component values to an accuracy of better than 5 per cent using only two controls and a unique solid-state null detector. Null-detector output is determined by two high-intensity LEDs. Design Mate 3 completely wired, tested, and calibrated, with technical data, is priced at $54.95.

Also available from Continental is a matching blank utility box. You can use it to house instruments of your own design whose appearance will match the Design Mate family. The utility box is the same size and shape as the Design Mate instruments. It's made of durable, high-impact, high-temperature plastic and is furnished with predrilled metal bottom plate and mounting hardware. The utility box sells for only $5.50.

For comprehensive specifications on the three Design Mate instruments, a 20-page illustrated catalog is yours for no charge. Write Continental Specialties Corporation, 44 Kendall Street, Box 1942, New Haven, Connecticut 06509, or Box 7809, San Francisco, California 94110, or use *check-off* on page 118.

125 Hz crystal filter for Drake R-4C

Sherwood Engineering has recently announced the availability of a crystal filter, 125 Hz wide at the 6 dB points, designed specifically for the Drake R-4C communications receiver. The new filter is completely compatible with the standard accessory filters offered by the R.L. Drake Company, and is being marketed as an adjunct to those with more standard bandwidths.

The new 8-pole crystal filter, the Sherwood model CF-125/8, has a 2.5 shape factor at the 6 and 60 dB points (bandwidth of about 325 Hz at -60dB), and exhibits less than 1 dB passband ripple. The input and output impedances are 50 ohms. Ultimate attenuation is greater than 100 dB. The 11 dB insertion loss of the CF-125/8 is similar to that of the Drake accessory filters.

For the CW operator who is looking for maximum selectivity, particularly during CW contests, this filter offers a significant improvement in receiving capabilities under adverse operating conditions. Due to careful design, the crystal filter does not display excessive ringing, even with strong signals. One well-known 160-meter operator, who used a CF-125/8 filter during a recent 160-meter CW contest, reported excellent results and concluded that it was one of the finest CW filters ever offered to the amateur community.

The new CF-125/8 carries a full money-back guarantee if you're not satisfied, and is priced at $125 from Sherwood Engineering, 1268 South Ogden Street, Denver, Colorado 80210.

Morse-code reader

The Atronics code reader is a compact, solid-state instrument that decodes Morse directly from your speaker and displays the resultant message in alphanumberic form on the front panel. A choice of readout size is available. The model CR-101 characters are 0.65 inch high by 0.42 inch wide (16.5 by 10.7 mm); model CR-101A characters are 0.2 inch high by 0.15 inch wide (5 by 3.8mm).

All characters, including punctuation, are displayed one at a time. Code speed, on-off, and audio level are set by front-panel controls. The speed control, with settings between 0-10, is used as an indicator only. For any setting, code speeds between 70-140 percent of that setting can be decoded. For example, if the code reader is set for 14 wpm, it will display any code speed between 10-20 wpm.

A light-emitting diode above the speed control indicates the expected length of a received dot. Another light-emitting diode above the level control indicates mark *(on)* and space *(off)*. The only connection required is a line between a phone jack on the code reader rear apron and your receiver speaker

More Details? CHECK—OFF Page 118

terminals. Input impedance of the code reader is 1000 ohms.

A radio teletype interface module (model TU-102) is available as an optional accessory. The TU-102 accepts 5-level code (start, five data bits, two stop bits). Teletype speed can be selected for 60, 75, or 100 wpm. Auto features CR, LF, FIG and letters are provided automatically.

The model CR-101 and CR-101A are priced at $225 and $195 respectively; the model TU-102 RTTY interface module is $85.00 (a $10.00 installation charge is made if the TU-102 is purchased separately). For more information write Atronics, P.O. Box 77, Escondido, California 92025, or use *check-off* on page 118.

hand-held scanning monitor

A hand-held scanner is a real convenience when you're walking around and want to keep on top of the action on the vhf and uhf bands. The Electra Company announces an addition to its product line called the Bearcat Hand-Held Scanners. Two models are available, a two-band version covering the low- and high-vhf bands, and a single-band version that covers uhf. Both models feature four-channel operation including LED channel indicators and individual channel lockout switches. Also included are an auto-manual selector switch and a volume and squelch control.

The Bearcats come equipped with a telescoping antenna, but provision for an optional loaded (rubberized) flexible short stub antenna has been included.

Weight is only 11 ounces (312g); size is 6¼ x 1½ x 2¾ inches (16x4x7cm). Four AA type dry cells are used for power. Sensitivity on vhf is 0.6 microvolt; on uhf, it's 1.2 microvolts. Audio output is 250 milliwatts. Scan rate is eight channels per second. Convenience features include a built-in belt clip and jacks for optional external antenna, earphone, battery charger, and ac adapter.

Both Bearcat models are priced at $129.95 (crystals not included). For more information, write The Electra Company, 300 South on East County Line Road, Cumberland, Indiana 46229, or use *check-off* on page 118.

automatic voice identifier

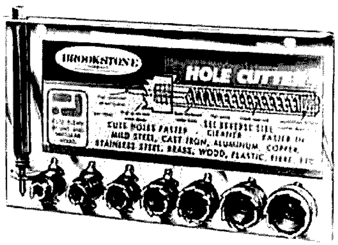
Newly introduced by Racom, Incorporated, is the Series 1500 voice identifier. Featuring high-reliability, all-solid-state circuitry, the Series 1500 uses the patented Racom disc principle (no tape loops) in conjunction with an electric timer that can be programmed in the field. No relays are used, although an option is available for those who might desire a "dry closure."

The Series 1500 can be field programmed to identify after each programmed time period, identify once after the last transmission, or identify after each transmission. Dual transmissions are prevented by a built-in channel monitor that can sense either audio or dc voltages. Operational status can be checked by illuminated indicators on the front panel. You can record any message, such as your dispatcher's voice. An erase interlock circuit ensures against accidental erasure of the recorded message. The Series 1500 automatic voice identifier can be mounted in a rack or on your desk.

Delivery from Racom is 4-6 weeks after receipt of order. For more information write Racom, Incorporated, 5504 State Road, Cleveland, Ohio 44134, or use *check-off* on page 118.

More Details? CHECK—OFF Page 118

miniature hole saws



There is an awkward range of hole sizes for electronic sheet metal and PC board material for which a series of miniature hole saws work very well. The Blair Equipment Company offers a set of seven hole saws in steps from one-quarter to five-eighths inch (6.5 - 16mm) with an interchangeable common arbor that obviates any need for drilling a pilot hole first. The arbor has a spring-loaded pilot which recesses into the arbor as the hole saw blade approaches the work, so you don't need a pilot hole for the arbor pilot to pass through. The blades, which are precision made, cut clean continuous chips through metals, shim stock, wood plastics, rubber, cardboard and other materials, leaving almost burrless holes. Each high-speed cutter is good for over 3000 holes in sheet steel and may be used with any ¼-inch (6.5mm) electric drill. The arbor has a high-speed steel automatic center point to avoid the need to center punch. a depth rod adjusts depth of cut up to 3/16 inch (5mm).

Blair specializes in automobile body equipment and markets these saws through stores catering to the auto body repair trade, but they are also available from Brookstone Company, Peterborough, New Hampshire 03458 for $20.95.

IC op amp cookbook

This new book by Walter Jung not only explains the basic theory of the IC op amp in a down-to-earth and easy to read manner, it also shows by example how to effectively use op amps in useful circuit applications. Fully illustrated, this practical book is bound to appeal not just to amateurs, but to anyone who has an interest in modern, linear design techniques — including amateurs, technical and engineering students, and practicing technicians and engineers.

The book is organized into three parts; Part I introduces the IC op amp and discusses general considerations, Part II covers practical circuit applications, and Part III consists of two appendixes of manufacturers' reference material.

Chapter 1 covers basic theory, which includes the ideal and nonideal op amp, with detailed analysis of error sources and dynamic characteristics. Chapter 2 describes early IC innovations, and discusses in detail the circuitry of such popular general-purpose types as the 709, 101, and 741.

Specialized units are also introduced and their general uses discussed. Chapter 3 covers general operating proceedures, such as nulling, frequency compensation, and protection against abuses and failures.

The remaining 5 chapters of the book discuss the application of op amps in such circuits as voltage and current regulators, precision rectifiers, limiters, comparators, logarithmic amplifiers, instrumentation amplifiers, analog multipliers, low-level preamps, active filters and equalization circuits, power booster stages, sine-wave oscillators, multivibrators, function generators, and voltage-controlled oscillators.

Unique IC op amps are also treated, such as programmable op amps, operational transconductance amplifiers, and quad current-differencing amplifiers.

This book is quite a departure from previous works of its subject and style. In terms of depth and content on applications of IC op amps, it is a virtual *tour-de-force*; 591 pages with over 250 circuit diagrams to pick from. And, unlike too many other applications handbooks, the circuits are clearly annotated, with the governing design equations given as well as the particular component values. Throughout the book emphasis is given to selecting the optimum IC for the job.

The book is not a textbook, nor is it a cookbook in the true sense of the word. It is really a "how to" cookbook that reaches the real-world level and approaches design problems as they actually occur. For this reason it is probably the single most valuable book available on IC op amps. If you now use op amps, or would like to, you'll find this book worthwhile. $12.95 from Ham Radio Books, Greenville, New Hampshire 03048.

More Details? CHECK–OFF Page 118

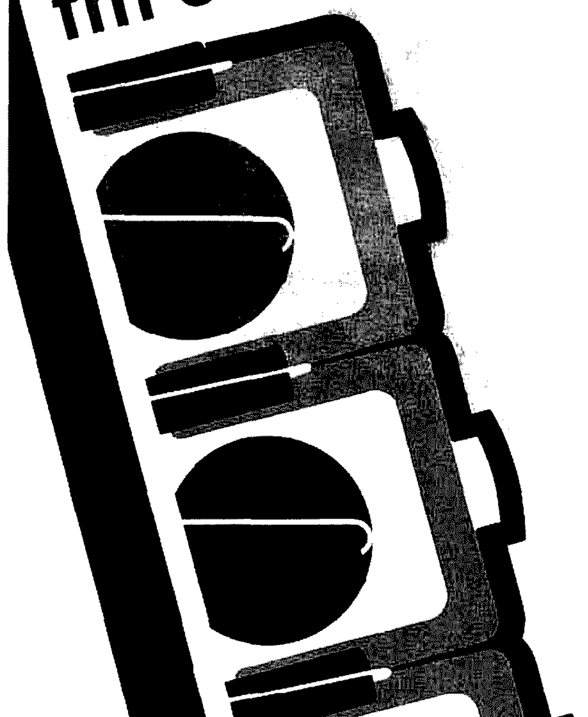
one dollar

ham radio

magazine

AUGUST 1976

high-performance two-meter fm exciter



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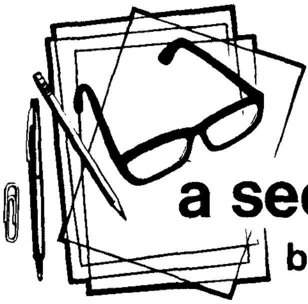


a second look

by Jim Fisk

> "At this very minute, with almost absolute certainty, radio waves sent forth by other intelligent civilizations are falling on earth. A telescope can be built that, pointed in the right place, and tuned to the right frequency, could discover these waves. Someday, from somewhere out among the stars, will come the answers to many of the oldest, most important, and most exciting questions mankind has asked."
>
> Frank D. Drake
> *Intelligent Life in Space*

In the late 1950s, soon after the United States became involved in a concentrated effort to place an astronaut into orbit around the earth, several scientific groups began to think seriously about using radio telescopes to search for extraterrestial intelligence. Inspired by articles in such prestigious magazines as *Scientific American* and *Sky and Telescope*, amateur radio astronomers began taking part in Project Ozma, pointing their antennas toward the heavens, looking for radio signals from intelligent beings in outer space. They didn't know where to look, nor what frequency to tune, so it's not surprising that when nothing was found for several years, enthusiasm began to wane. Then an English radio astronomer discovered the first pulsar, many scientists postulated that it might be a controlled radio beam from intelligent life in another galaxy, and a whole new search for extraterrestial communications was soon underway.

Unfortunately, there was no organized follow-up to Project Ozma, so it's not known today how many individuals or groups are still listening, nor who and where they are. Nick Marshall, W6OLO, a member of the very first Oscar group back in 1960, became interested in this problem and announced the formation of a "Reviva Ozma" committee at the 1975 Project Oscar meeting at Foothill College in California. One of the goals of Reviva Ozma (now known as Starquest) would be to locate and communicate with individuals and groups who were still listening, and to help them publish their findings. Another goal was to assemble a state-of-the-art listening post which would be dedicated to continuing the search for radio signals of extraterrestial origin.

In late 1975 Marshall contacted Dr. John Billingham at the NASA-Ames Research Center for further information on the Project Cyclops follow-on, a NASA study of a system for detecting intelligent extraterrestial life. It became immediately apparent that Starquest could serve an interim purpose by disseminating correct information about Project Cyclops and, more importantly, assist in trying to keep various listening frequencies clear of occupancy by commercial, industrial, and government radio services.

Marshall also made contact with members of Operation Serendip, a study being conducted by a small group from the University of California at Berkeley. This group has been using an 82-foot (25m) dish near Mt. Lassen to listen and record extraterrestial signals on 1420 MHz and process the data on a pdp-8 computer.

Foothill College graciously offered the Starquest group an antenna location near their observatory, and work is now underway on a spherical dish which is patterned after the 1000-foot (305m) dish at Arecibo, Puerto Rico, although on a much smaller scale. The initial Starquest dish will have a 35- to 50-foot (10.7m-15.2m) center section which can be expended later to about 100 feet (30m) in diameter. Starquest members are also formulating plans for a state-of-the-art receiving and data-processing system which will be used in the station. In the near future the group is expected to start publishing a quarterly *Starquest Bulletin* which will be mailed to members and affiliates.

In addition to their other activities, Starquest has established a world-wide amateur-radio net on 20 meters for gathering and disseminating information relating to the search for extraterrestial signals. This net, which meets on 14.280 MHz (±5 kHz) at 1900 GMT, the first Sunday of every month, is now operational. It is hoped that interested amateurs and experimenters will listen in, and help Starquest locate other interested groups in their own local areas who are involved in the search for intelligence in outer space. Amateurs living in Northern California who would like to participate directly in the construction of the Starquest listening station are invited to contact W6OLO for more information.

Jim Fisk, W1DTY
editor-in-chief

presstop

FCC'S FIRST ACTION ON RESTRUCTURING was taken in mid June, and Novices and Technicians became the principal beneficiaries. This first Report and Order on Docket 20282 liberalizes both license classes' privileges but appears to reject the "Dual Ladder" approach to restructuring that was such an integral part of the Docket's original philosophy.

Technician Class Licensees receive the most significant benefit in the form of full Novice privileges — CW on 80, 40, 15 and 10 meters. No other Technician privileges were changed at this time, so it's still possible that 10-meter phone and/or expansion of Technician frequencies on the bottom of two or six meters may be part of expected later actions on 20282. Technician exams will no longer be "mail order," either — the Report and Order makes administration of Tech exams an FCC Field Office job.

Novices Receive two benefits under the just announced Rules changes. Power limits for Novices have been raised to 250 watts, and a Novice who has failed to upgrade during his two-year license term will be permitted to continue as a Novice by taking another Novice exam — the requirement for a one-year delay before going for another Novice license has been dropped. The new 250 watt Novice power limit actually applies to Novice frequencies, by the way — any station operating in the Novice sub bands will be limited to 250 watts, regardless of license class.

All "Mail Order" Amateur Exams except Novice are effectively being eliminated. The one exception that will remain is for an applicant who cannot travel because of medical reasons and who can back it up with a physician's statement.

Deletion Of Conditional Amateur Licenses won't shut the door on Americans overseas such as servicemen and missionaries who can't get to an FCC Field Office for their exams. Procedures to take care of such people are being developed and they will be taken care of on a case-by-case basis. One idea being considered is a "Conditional" license good only while the holder remains out of the country — he'd have to be reexamined by the FCC on return if he wished to continue his Amateur activities.

All Present Conditional and Technician (C) license holders will be "grandfathered" into straight Technician and General Class licenses as their present licenses come up for renewal. No re-examinations will be required, which should make at least a few present license holders breathe easier!

AMATEUR LICENSING TIMES have definitely dropped since our last report, with several recent renewals returned 37 to 45 days after they went into the mail. As of early July the FCC was handling requests for volunteer exams in about two weeks, down from six weeks in early June.

The Next Big Hurdle is the phase-in of WD calls, where computer programming problems have developed. Until they're overcome, significant delays in getting new licenses out are likely to begin to occur despite efforts to head the problems off.

AMATEUR RADIO IS VERY PROMINENT in the Smithsonian Institution's newest displays. OSCAR I is on display in the Communications Satellite area of the new Hall of Satellites, and NN3SI, the Institution's new Amateur station located in the Museum of History and Technology, was dedicated on June 8th. First contact from NN3SI was on CW with trustee W4KFC working W1AW with the same key General Sarnoff used in 1912 when he participated in the Titanic disaster.

NN3SI Is the Institution's special events call — WB3APS is its regular call — and NN3SI has been authorized for a year with operation on all bands and OSCAR planned.

Smithsonian's New Amateur Station can be a bit tricky to get to, warns K8NHR. It's almost directly under the antennas, and can be found by bearing hard left after entering the building's main entrance.

UNLICENSED 27 MHZ OPERATORS are protected by a legal loophole — the Communications Act of 1934 gives FCC jurisdiction only over licensees. Justice Department is the agency with the power to go after unlicensed operators, but it has neither the tools (as FCC does) nor time to chase down all such violators.

This "Catch 22" May Change shortly, however — FCC Chairman Wiley said at the CES show in Chicago in mid June that there is legislation pending before both houses of Congress that would give the FCC power to prosecute unlicensed operators and confiscate their equipment, and he expects it to become law very quickly.

TWO-LETTER CALLSIGN availability chart has been prepared by the ARRL from current FCC records and can be had for the asking. It covers all ten districts and includes instructions on the procedure for requesting a two-letter call. Send an SASE with 24¢ postage to ARRL for a copy.

MOONBOUNCE ENTHUSIASTS are going to get a crack at Alaska thanks to K6YNB/KL7. Wayne plans to open up from Ketchikan, August 11th on 144 and switch to 432 about the 18th. He also plans to be on 50 and 220 MHz, but not with the same big gun antennas that he'll use on the other bands. Meteor scatter (the Perseids shower occurs during that period) as well as possible tropo and Aurora contacts are also expected during the trip.

high-performance two-meter fm exciter

Eleven channels, superior modulation and complete kit availability are just a few features of this little jewel

After reviewing **the limited number** of two-meter fm transmitter construction articles available to the homebrew enthusiast, I decided it was time to break away from the Sonobuoy-type design and try to generate some interest in building a more conventional commercial-type exciter. This article is the result of the overwhelming response to an earlier construction article for an fm receiver[1] of the type of design I am encouraging.

Before dismissing the Sonobuoy-type exciter completely, I'd like to mention that these designs, which have appeared in the amateur literature for the past few years, deserve a great deal of credit. They were easy to build and were the first solid-state transmitters to gain wide popularity and to be constructed in quantity. However, they had some disadvantages that I've attempted to correct:

1. The Sonobuoy designs used direct frequency modulation of the oscillator with a varicap diode. Although sometimes described as a feature such modulation often resulted in unsymmetrical modulation because of improper dc biasing. For some reason, this flaw has been perpetuated in several spinoffs of the Sonobuoy transmitter that I've seen.

2. Direct fm circuits made crystal switching difficult. Such circuits could not be used with a frequency synthesizer.

3. Audio circuits weren't really optimized for voice operation with a variety of microphones; not surprising since Sonobuoys were designed for a different purpose.

4. Tuned circuits were unshielded and construction, in general, was intended to be of the "disposable" type

By Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14162

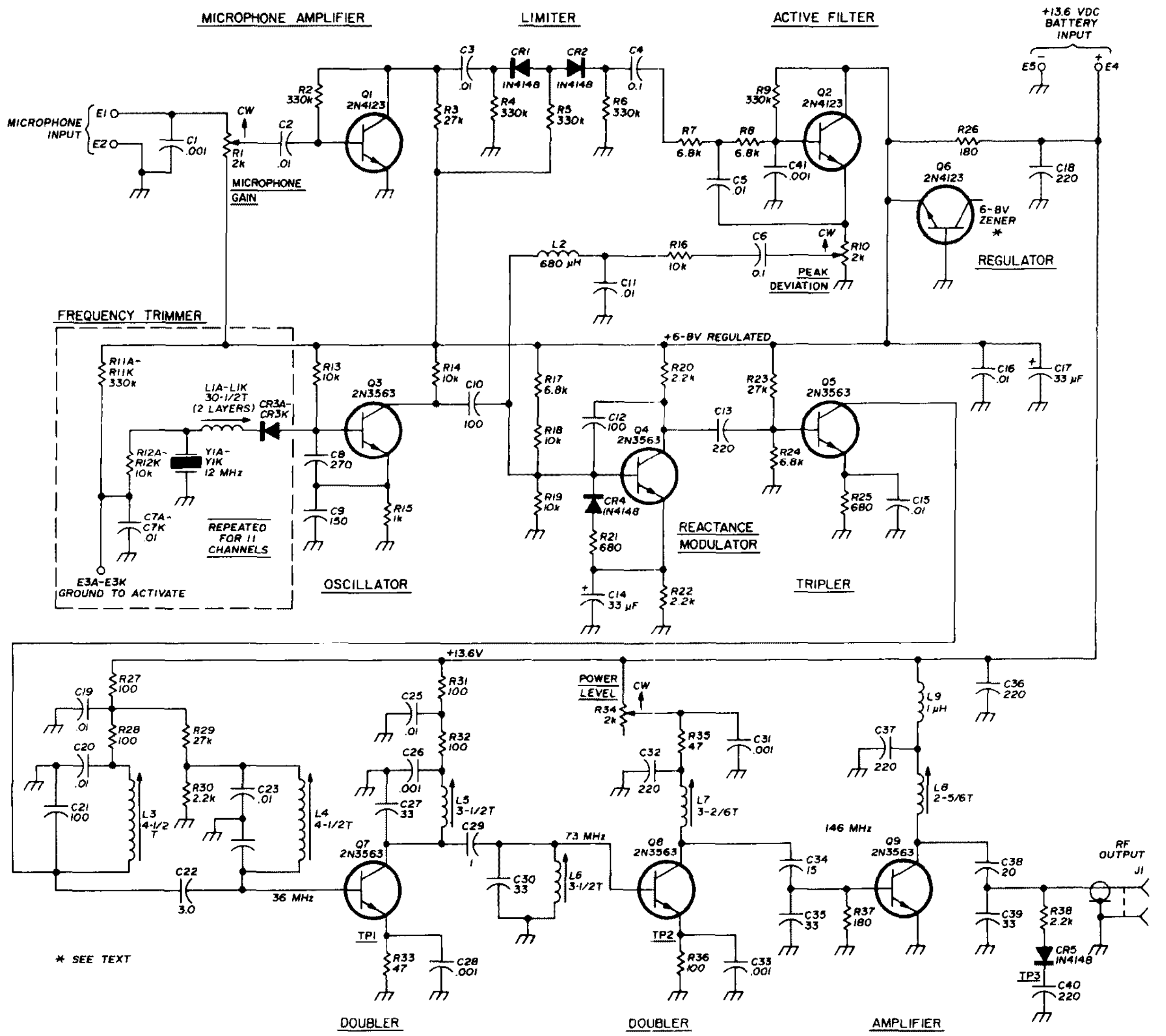


fig. 1. Schematic of the high-performance 2-meter fm exciter. A complete parts kit is available together with other accessories at modest cost.

because of the original intended design function. You can readily see the difference when such a unit sits next to a commercially designed two-way radio.

design goals

After sorting through the many available circuits for a two-meter fm transmitter, design objectives were based on the best features of some of these circuits that could be implemented with readily available parts and simple construction.* The design goals were:

1. Superior modulation – quality to be proud of in either a repeater or home station.

2. Inclusion of both deviation and microphone gain controls.

3. Effective lowpass filter following the modulation limiter to eliminate raspy voice signals.

4. Compatibility with either carbon or transistor-amplified dynamic microphones.

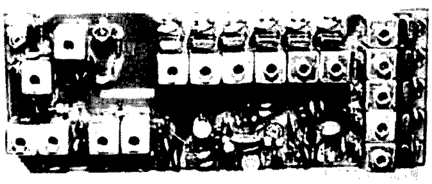
Here's an example of *ham radio's* promise to bring you construction articles based on available kits so that parts scrounging will be a thing of the past. This two-meter fm exciter has been designed with the homebrewer in mind – easy to build and easy to align and get working. You'll find many features that are included in the latest commercial designs. Best of all, the parts kit price won't fracture your pocketbook. **Editor**

*A list of kits and accessories for the transmitter will be found at the end of this article.

5. Phase modulation suitable for use with multichannel operation and frequency synthesizers.

6. Shielded coils and sufficient tuned circuits between multiplier stages to reject harmonics and spurious signals.

7. A maximum number of electronically switched and individually adjustable channels, consistent with the selected board size along with exciter and multiplier stages.



Top view of the completed 2-meter fm exciter showing clean layout of parts without overcrowding.

8. An oscillator using common 12-MHz series-resonant crystals.

9. Voltage regulator for oscillator, modulator, and audio stages to minimize effects of line-voltage variation and noise.

10. Adjustable output level.

11. Sufficient output power (150-200 milliwatts) to drive the new rf power modules now available.

12. A design easy to build, align, and test.

features and technical characteristics

The exciter measures 3 x 7½ x 1 inches (7.6 x 19 x 2.5cm) and weighs 5 ounces (142gm). Operating power is +13 volts at 70 mA. Sufficient power is provided for use as a control link, or the exciter can be used as a low-power transmitter. It can be adapted easily to 50 or 220 MHz by changing the multiplier tuned circuits, or it can be adapted to 450 MHz by adding tripler/driver stages on a separate board.

Eleven oscillator channels are used. Multilayer oscillator frequency-netting coils provide vernier adjustment on each channel. A simple ground-*on* diode channel-switching scheme allows easy adaptation to trunk mounting (important in these days of ripoffs) or other remote-control application. CW operation is possible by keying the B+ line to the multiplier and output-amplifier stages.

The exciter can also be used as an inexpensive multichannel frequency standard or fm signal generator. A convenient output-level control, used with an external fixed attenuator, provides variable output level.

The exciter schematic is shown in **fig. 1.** Audio signals from a carbon or transistor-amplified dynamic microphone are applied to microphone amplifier Q1 through microphone gain control R1. This control, normally not provided in most transmitters, allows the audio level to the limiter stage to be adjusted independently of deviation level, so that sufficient audio punch is obtained without excessive clipping. Limiter CR1, CR2 consists of back-to-back diodes forward biased a small amount. When audio peaks exceeding the bias level are applied, clipping action limits the audio transferred to the next stage. Q2 is a lowpass filter with a cutoff frequency just above the normal voice range to remove the audio harmonic components generated by clipping the audio signal. This stage, together with its separate microphone gain and deviation controls, is primarily responsible for the professional-sounding modulation of this exciter.

The 12-MHz injection to the modulator is provided by Clapp oscillator Q3. Eleven channels are diode switched by grounding the appropriate control lines. The diodes are reverse biased except the one having its control line activated to cause dc conduction, thereby completing the path to one crystal circuit from the base of Q3. The variable coil allows the crystal load reactance to be varied for frequency netting.

The phase-modulated 12-MHz signal from reactance-modulator stage Q4 is multiplied in tripler stage Q5, doubler Q7, and doubler Q8. Double-tuned circuits between multiplier stages provide rejection of spurious frequencies. (Multipliers create more than one harmonic, of course; and if tuned circuits of sufficiently high Q are not used, undesired frequencies will be passed through the multiplier chains to cause spurs in the transmitter output).

The B+ voltage to doubler Q8 is adjustable with series potentiometer R34 to provide output power level control. This control is normally set fully clockwise but may be used to reduce output level if desired. (For example, to limit drive to a power amplifier or provide variable output when used as an alignment generator). Amplifier Q9 provides 150-200 mW output to a 50-ohm load (2-3 volts). This level was chosen to drive an rf power module and is also suitable for several other applications. If desired, a simple one-watt PA stage could be added as well as many other types of amplifiers, although the rf power module is simplest by far.

To allow for easy alignment with only a vtvm, three test points provide dc signals as a function of rf levels at several stages. TP1 and TP2 provide indications of emitter current in the two doubler stages. TP3, in conjunction with rf detector CR5, C40, provides an indication of rf output to the antenna or power amplifier.

power amplifiers

There is only one word to describe the new rf power modules by TRW and Motorola: "fantastic." You have probably seen them in advertisements for various radios.

The photo of a power amplifier using one of the rf power modules shows how simple it is to make a PA of moderate power level today. The exciter in this article was designed to drive these power-module PAs.

The rf power modules, or "bricks" as they're sometimes called, are magic compared to the alternative. Each PA brick is an integrated circuit containing several power amplifier stages with decoupling and tuned circuits to provide many watts output for 150-200 mW input. All that's required externally is a means of connection (a PC board) and low-frequency decoupling components.

Operated at normal 13.6 volts from a battery supply, the PA bricks provide 20 watts of rf at 145 MHz, or 13 watts at 432-450 MHz. (A tripler/driver module kit, model T-20, is available for 450-MHz operation). At this power level, no damage will occur even when operating into a load of infinite vswr. Usually, the bricks can be driven to 25 watts at 2 meters or 15 watts at 450 MHz or higher if loaded properly and operated with sufficient drive and B+ supply. Of course, the greatest feature is that absolutely no tuning is ever required! Current requirement is 2-4 amperes, depending on rf level. Efficiency is 30 to 50 per cent.

No doubt it's more expensive, watt-for-watt, to use an rf power module instead of an amplifier with discrete components. However, considering the difficulties you can have with discrete PAs and the careful tuning and special packaging required, the brick is a bargain and certainly adds a lot of pleasure to homebrewing transmitters.

construction

The exciter is assembled on a single-sided 3 x 7½ inch (7.6x19cm) PC board. Construction details are shown in fig. 2. The following details of coil assembly and other suggestions are given to facilitate assembly.

Plastic coil forms of ¼ inch (6mm) OD are used with ½ inch (13mm) square shields and vhf tuning slugs. The coils (fig. 1) are wound in a clockwise direction as viewed from the top, using the solderable wire supplied with the kit. *All turns are close spaced.* The diagram of fig. 3 is exaggerated for clarity. However, all leads should be pulled tight, no fancy bends are required, and no Q dope is needed. Holes in the base of the form are numbered as shown. Fractions of turns relate to the number of funnels on the form; for example, 2/6 turns beyond a full number (as in L7, fig. 1) means ending two funnels *beyond* the starting funnel.

Oscillator trimmer coils are wound with no. 26 (0.3mm) wire in two layers, with just over half the turns on the first layer and the remaining turns on the second layer. Wind from bottom to top, going directly back to the bottom and winding up again to the top on the second layer. All other coils are wound in one layer with no. 24 (0.5mm) wire. Coils with capacitors will have capacitor leads inserted through coil-form funnels after the coils are wound. The capacitors should be seated down onto the board as much as possible, as should all parts. This is important at vhf!

Don't be over concerned with coil winding. Neatness isn't important; turns can overlap and windings need not

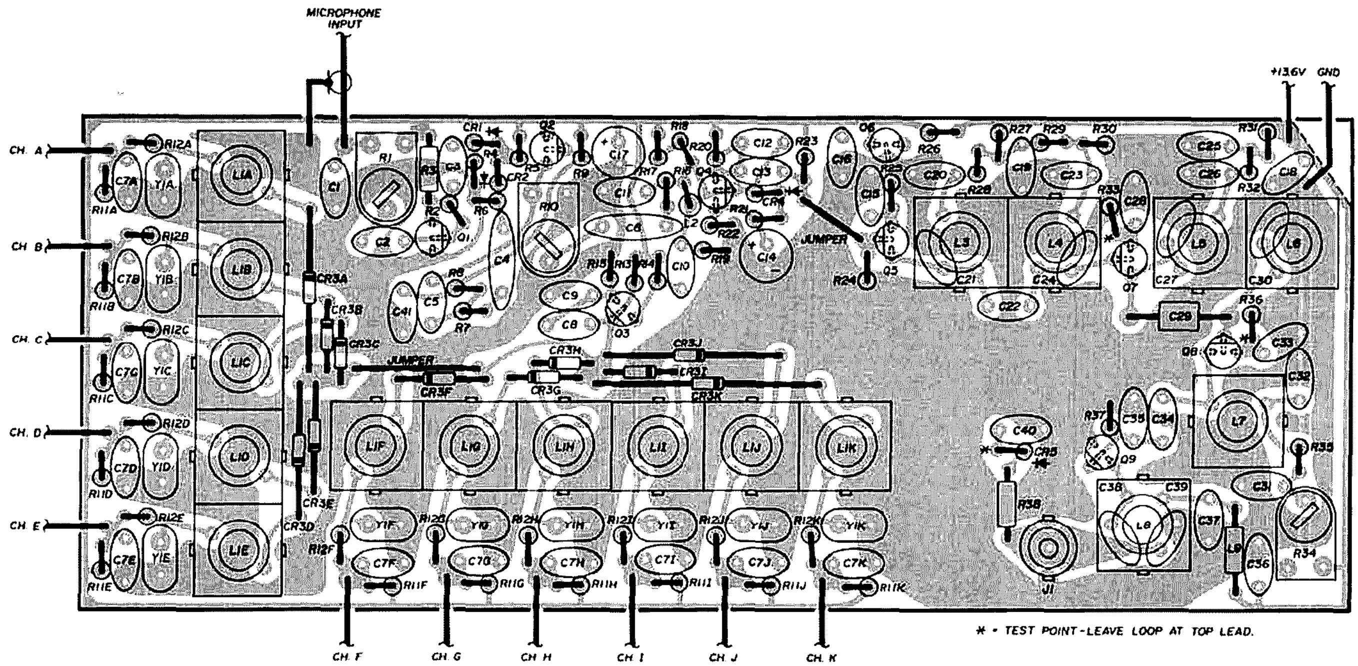
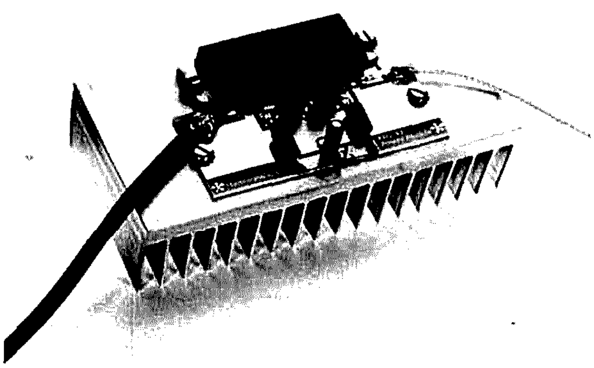


fig. 2. PC-board component layout for the 2-meter fm exciter. Eleven channels are included plus microphone gain and deviation-level controls all on a 3 x 7½ inch (7.6x19cm) printed-circuit board.

be uniform. Start coil leads through board holes while the coil form is above the board, then seat the coil into place onto the board. Don't attempt to insert capacitor leads with the coil form tight against board. After coils are installed, apply heat from a very hot soldering iron



The 2-meter fm exciter married to one of the "instant PAs" for increased power output.

for 10-15 seconds with solder applied. This will automatically strip the wire and allow solder bonding to occur. If you prefer, the leads may be stripped in the conventional way before installation in the board. Don't solder-strip the leads unless the coil is mounted on the board, or the leads will migrate in the warm plastic.

Shield-can tabs should be soldered to the board for proper grounding and mechanical support. It is unnecessary to bend the tabs. Diodes and electrolytic capacitors must be installed with proper polarity. The top leads of R33, R36, and CR5 should be installed with loops accessable for probe connection. The extra length allows the top of the component to serve as a test point. Crystal sockets may be used with the popular HC-25/U type crystal, or crystals may be soldered directly to the board.

External leads should be connected by soldering them into large pads on the PC board. The microphone shield may be easily terminated by wrapping a few turns of bus wire around the shield and soldering. The channel switch should ground the desired channel line for activation. It's sometimes possible to use a common switch for both transmit- and receive-channel selection if the receive switching scheme also uses a ground.

mounting

The T40 PC board is designed to slide into a vertical groove in a companion cabinet just forward of the rf power module mounted on the rear panel. The components should face forward, and the board should be oriented upward as shown in the component location diagram. The upper right-hand corner may be cut off as indicated in **fig.** 2 to allow cables to be routed past the board in the cabinet. If the companion cabinet is not used, appropriate holes may be drilled in ground areas of the PC board to allow standoff mounting to your own chassis or panel.

crystals

The exciter uses HC-25/U series-resonant crystals. The crystal frequency, in the 12-MHz region, is determined by dividing the 2-meter channel frequency by 12. When adapted for other bands, the divisor changes accordingly; e.g., 36 for 450 MHz, 18 for 220 MHz, and 4 for 52 MHz.

alignment

After constructing and visually checking the PC board for proper assembly and soldering, you are ready to apply power and perform alignment and testing. Caution: Use a proper tuning tool; a loosely fitting tool may crack the powdered-iron tuning slugs.

1. Install one crystal at the approximate center of the desired frequency range, and ground the corresponding channel control line.

2. Set the audio controls R1 and R10 to full counter clockwise, and set power control R34 fully clockwise.

3. Preset all tuning slugs to half range.

4. Connect 50-ohm load to output connector J1.

5. Apply 13.6 Vdc (battery power, etc.). *Observe polarity.*

6. Check regulated voltage. It should be approximately +6 to +9 Vdc.

7. Connect vtvm, set to 0.5 Vdc range, to TP1 (loop at top of R33 at first doubler Q7).

8. Peak L3; then peak L4. Dip L5. Dc voltage should be roughly +0.3 to +0.4 volt.

9. Connect vtvm, set to 1.5 Vdc range, to TP2 (top of R36 at second doubler Q8).

10. Peak L6; then peak L5. Dip L7. Dc voltage should be roughly +1 to +2 volts.

11. Connect vtvm, set to 1.5 Vdc range, to TP3 (top of CR5 at final amplifier Q9).

12. Peak L8; then peak L7. Dc voltage should be roughly +1 to +2 volts.

13. Repeat all above steps to eliminate effects of interaction and to check tuning. When you're finished, the dc current drain should be about 70 mA. If you have an accurate rf probe arrangement, the rf voltage at the 50-ohm load should be about 2-3 volts (150-200mW), which is the level required to drive the rf power module. Note that output will be somewhat less if the power supply voltage is below 13.6 Vdc. *Note*: Do not attempt to tune in any manner other than described. In particular, multiplier stages should *not* be repeaked for maximum at the antenna connector.

14. One, by one, ground each channel control line with crystals installed, and adjust the corresponding oscillator trimmer coil to net each channel to the proper frequency.

troubleshooting

The usual troubleshooting techniques of checking dc voltages at transistor elements and tracing ac signals, with a voltmeter and an rf probe where applicable, apply in this case. Don't overlook the possibility that parts may be installed incorrectly.

For convenience, the regulated voltage is obtained for the low-level stages by using a 2N4123 as a zener diode (Q6 in **fig. 1**). Since transistors are not calibrated for this parameter, the zener voltage should be checked the first time the board is fired up to ensure that a zener voltage in the range of 6 to 9 Vdc occurs. If lower, you may wish to substitute another 2N4123 to find one with a useful zener voltage. The exact voltage is unimportant; it is only necessary that the voltage be held stable under varying line conditions.

If trouble is encountered in netting one or more channels, check the number of turns on the corresponding oscillator trimmer coils. Make sure the coil is wound in two layers, as described earlier. If a channel won't oscillate at all, check the corresponding diode and other components in the control-line circuit. The following typical test voltages will serve as a *rough guide* to proper transistor operation, based on 13.6 Vdc input.

transistor	emitter	base	collector
Q1	0	0.6	2
Q2	2.2	2.9	7*
Q3	3.8	4.4	6*
Q4	2.6	3.2	4.5*
Q5	1.2	1.8	13.4
Q6	7*	0	-------
Q7	0.35†	0.6†	13.4
Q8	1.7†	1.7†	13.6

*Assumes 7 volts from a regulated supply, but supply ranges between 6-9 V in actual units.

†Rough indication of drive level.

Base and collector voltages of Q9 cannot be measured with drive applied because of rf effects on meter.

microphone and audio adjustments

The exciter is designed to operate with either a carbon microphone or a transistorized dynamic microphone. The microphone should be connected with shielded cable to avoid rf pickup. To adjust deviation level, preset R1 and R10 both fully clockwise. Key the exciter, and make sure that the carrier is adjusted properly to frequency. Speak into the microphone and observe the deviation meter on the receiver, or listen to the audio with the squelch set tight. Reduce deviation control R10 setting until all noticeable effects of overdeviation are removed; e.g., distortion, meter swing on peaks, squelch pumping.

The setting of microphone gain control R1 is a refinement not found on most transmitters. It should be set to provide sufficient audio for full modulation on voice peaks but low enough to remove background noise and obvious clipping effects, which normally result from overdriving a clipper.

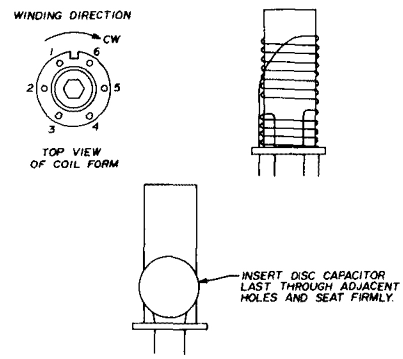


fig. 3 Diagram for winding coils (exaggerated for clarity). The number of turns for all coils is shown in the schematic. All turns are close spaced; winding direction is extremely important.

Since the amount of deviation depends on the frequency band to which the exciter signal is multiplied, resistor R16 is provided to allow the amount of deviation at full control setting to be altered. The value of R16 may be reduced to obtain wider deviation if required, and it may be increased to reduce the range of the deviation control. (R16 is also a part of the lowpass filter following the limiter).

reference

1. G. Francis Vogt, WA2GCF, "High Performance VHF FM Receiver," *ham radio*, November, 1975, page 8.

ham radio

update of the phase-locked loop RTTY demodulator

Here are the answers to your questions about this terminal unit plus a modified circuit for upward shift

Since the NS-1 phase-locked loop RTTY demodulator first appeared in the *RTTY Journal*, October, 1974, and later in *ham radio*,[1] I have received numerous inquiries about its operation. I hope this article will answer some of the questions and also provide some added tips.

First, as an explanation, the NS-1 was developed primarily for fsk downward shift on the high-frequency bands, using low tones to take advantage of a narrow receiver passband. The 741 op-amp limiter drops off around 2000 Hz, so the tones must be within this limit. On fsk the tones can be varied by receiver tuning, so it's easy to obtain these low tones. Shift reversal is accomplished by changing sidebands if receiving in the ssb mode or by moving the bfo to the other side of zero beat if receiving in the CW mode.

meter adjustment

I suppose the most-often asked question concerned the value of the zero-center, meter-adjustment pot. This value depends on the meter movement; i.e., whether it's 50 μA, 200 μa, 1 mA, etc. The easiest way to determine this value is to put a 500k pot in series with the meter and gradually reduce the resistance until full scale (+) is obtained on the mark/hold signal. This resistance can then be measured and a fixed resistor substituted.

If you don't have a zero-center meter, fig. 1 shows such a circuit using a regular 0-1 mA meter. Before connecting the 18k resistor to the *M* terminal, set the 25k pot about midway, ground the 18k resistor to the chassis, and adjust the pot until the meter reads center scale (0.5mA). Now, when the 18k resistor is connected to the *M* terminal, the meter will act like a zero-center meter with plus to the right and minus to the left.

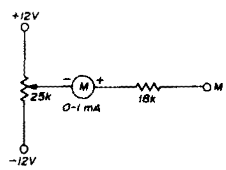


fig. 1. Using a 0-1 mA meter for zero-center indication. Connected as shown, the meter will indicate plus to the right and minus to the left. See text for adjustment instructions.

cross pattern

Several readers asked where to connect a scope to receive a cross pattern. A scope cross *pattern requires* tuned filters to distinguish between the mark and space tones, one displayed vertically and the other horizontally. Since the NS-1 has no filters, there is no place to connect the scope. However, if you have a scope with tuned filters that *will* produce a cross pattern, connect the scope ahead of the NS-1 or at the receiver output. Tune in a signal that gives a good cross pattern, then adjust R1, the 5k vco pot, until you get good copy. Thereafter you can use the cross pattern for tuning.

Some readers complained that they were unable to

By **Nat Stinnette, W4AYV**, P.O. Box 1043, Tavares, Florida 32778

get wide-shift copy. The usual 2125/2975 Hz tones will not work, as the 741 limiter will not accept frequencies much above 2000 Hz, as stated previously. So tones, say, around 2125/1275 Hz should be used. A compromise setting of R1 can be found that will permit copy of both wide and narrow shifts.

other tips

The purpose of the switch is to put the teletype machine in a "hold" condition. Random noise and signals will produce garble when tuning. Also the switch should be closed when transmitting, as interaction between the receiver and TU will also produce garble.

Remember that the NS-1 will act very much like any other fm demodulator. If a stronger signal is near the one you're trying to copy, the stronger signal will take over. I've found that a receiver T-notch filter is quite effective at times. Also, use the narrowest selectivity you have on your receiver that will permit copy of the tones; this reduces the effects of interference. One amateur wrote that he installed an active audio filter (2100-2300 Hz) ahead of the NS-1 and adjusted the unit for 2295/2125 Hz. He said this arrangement compared very favorably with the ST-type terminal units.

modified unit for upward shift

A few people have tried to use the NS-1 on afsk on the six- and two-meter bands and found it would not work. This is because afsk is usually upward shift (mark low, space high), and the NS-1 will not copy this way. Afsk tones are fixed. Nothing can be done at the receiving end to reverse the shift, and there are no provisions for it on the TU.

To achieve copy with upward shift, it's necessary to reverse the polarity of the voltage coming out of the 741 comparator on the mark/hold signal. Several circuit changes were tried including reversal of the two inputs to the 741. This change worked, but best results were obtained after installing a transistor just ahead of the 2N5655. A 2N706 was used. It switches the keying transistor off and on and has the effect of reversing the voltage from the 741 output, which permits smooth upward-shift copy.

The 741 limiter was eliminated since it is restricted to around 2000 Hz; thus high tones such as 2125/2975 Hz can be used. Two reversed diodes were placed ahead of the 565 PLL. These give good limiting and prevent PLL overload. The 565 is very sensitive, so no amplifier is needed. This modified arrangement works equally well on afsk and fsk, wide or narrow shift. The revised circuit diagram is shown in **fig. 2** and has been designated the NS-1A.* The same adjustment procedure as used on the NS-1 should be used for the NS-1A. The zero-center tuning meter will now show full scale minus (left) reading on the mark/hold signal. The tuning meter may not even be needed on afsk.

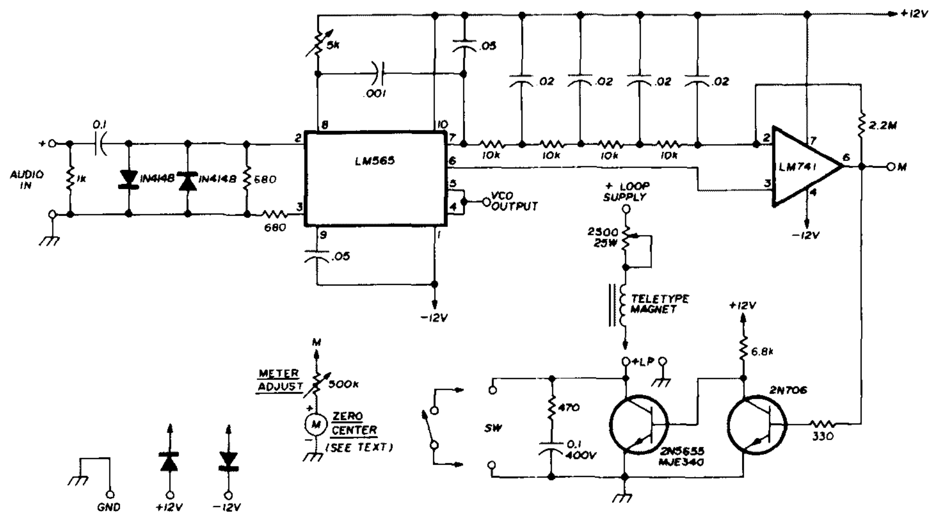


fig. 2. Schematic of the NS-1A phase-locked loop demodulator for copying afsk/fsk (upward shift). The 2N706 switches the 2N5655 on and off, which reverses the polarity of the voltage from the LM741 comparator on mark/hold. This permits smooth upward-shift copy.

acknowledgements

I wish to thank Ron, W8BBB; Buck, WAØLEM; and Rick, WB5FHU for their tests and evaluation of the NS-1A on afsk. The added feature of afsk copy should encourage more activity on two- and six-meter RTTY with this simple terminal unit.

*At present no modified boards are available; however, wired and tested NS-1A units are available for $29.95 postpaid; the circuit board is $4.75 postpaid. For more information, contact the author.

reference

1. Nat Stinnette, W4AYV, "Phase-Locked Loop RTTY Terminal Unit," *ham radio*, February, 1975, page 36.

ham radio

the hand-held electronic calculator: its function and use

First of a four-part series on how to apply the simple four-function machine to math problems encountered in radio work

This is the first of four articles on the hand-held electronic calculator — how it works and how it can work for you to solve even the most complex problems encountered in radio work. The operating principles of the four-function machine are discussed first, with examples illustrating how the basic arithmetic functions are performed in response to input-key manipulations. Discussed next are some of the more important operations that can be performed with larger and more expensive machines: chain operation, the use of constants, and the use of various memories. Finally, some suggestions are given to help you choose the best machine to fit your needs.

In the second article, which will be published in the next issue of *ham radio,* the power of the simple four-function calculator is expanded for solving problems in radio work. Examples are given on using approximations as substitutes for special calculator functions, the use of problem-organization technique to overcome the limitations of simple machines, and the use of the "scratch pad" as a substitute for calculator memory. Examples are given on using the simple four-function machine to solve problems involving transcendental functions and numerical integration. The following articles will cover transmission-line calculations and use of the new programmable calculators.

These articles should help dispel the mystery of how to use the four-function electronic calculator in solving electronics problems by providing basic information on the logic of operations within the machine and providing the approximations and tools available.

basic calculator principles

Addition. In digital electronics, a *register* is a place to *store* numbers. Actually, these don't have to be electronic devices; a chain of relays, a toothed wheel, or even knots in a piece of string can serve as a register. But in the small calculator registers are always electronic and are almost always made using field-effect transistors.

To be useful there must be a way of putting numbers into the register, of taking them out, and, for at least some registers, of telling what number is stored. Hand-held calculators have standardized on a set of push-buttons or *keys* for *input,* and a stylized *display* for presenting the content or number stored.

Suppose we wish to perform the simple arithmetic operation of addition. We could use three registers, one for the *addend,* or first number, a second for the *augend,* or second number, and a third for the result, or *sum.* Actually it's easier to do this with only two registers by a technique called "add to storage." In this, the addend, when received, is first placed in the sum register, then the augend is added to it to get the final *sum.* The two

By R.P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida 32019

registers involved can be called the keyboard register and the answer register.

To make a small calculator, we need the two registers, the keyboard, and the display. We also need some control circuitry to tell the register how to work. For convenience, the control can also be arranged to switch the display from the keyboard to the answer register. In block diagram form, these elements are shown in **fig. 1**.

We can see these registers in operation on any calculator by considering only the ten number input keys 0 through 9, and the *add* key. This key is usually marked +/= in small calculators, but may be marked +. When the calculator is first turned on, a *0* will appear at the extreme right of the display (a decimal point may appear also, but ignore this for now). Suppose we wish the addend to be the number *123*. Pressing the *1* key causes the *0* to change to a *1*. Pressing the *2* key causes the *1* to jump, or shift, to the second position from the right and a *2* to appear at the rightmost position. Similarly, pressing the *3* key causes the numbers displayed to shift again, the display now reading *123*. We have now *entered* the desired number, *123*.

To transfer this number to the answer register, we now press the add key (+/= or +). All that seems to happen is a blink of the display, but the display is now presenting the number in the answer register. We can show this as follows. Press the *0* key, which causes the display to change to a zero. Now press the add key again: the display changes to *123*, the number stored in the answer register. This display will continue to show the number in the answer register until another operation is started.

In this test we are actually performing an addition, *123 + 0 = 123*. We can continue addition by entering another number then pressing the add key. Each time, the display switches to the keyboard register then back to the answer register as the add key is pressed. The answer register always shows the sum of all numbers entered.

When pressing the number keys it's possible to make a mistake, say by pressing *132* instead of the desired *123*. This is easily seen, since the display shows the number actually entered; this is the reason for switching the display back and forth.

To correct a mistake in entry, it's necessary to wipe out the entire number entered, or *clear* the keyboard register, and start over. On most small calculators this is done by pressing a special control key marked *CE* for *Clear Entry*. In this example, pressing *CE* causes the display (and the keyboard register) to change from *132* to *0*. The correct number is then entered.

Let's leave addition and go on to other operations. To get ready for these, the number stored in the answer register must be removed lest it make an error in the next calculation. To do this, another control key is pressed, marked *C* for *Clear*, which sets the internal register to zero. The same result could be obtained by switching off the calculator; the internal circuits in most calculators are arranged to clear the registers when the power is first switched on.

In some calculators *Clear Entry* and *Clear* are combined into a single key, usually marked *C* or *CLR*. Pressing this key once clears the entry, and pressing it twice in succession clears all registers.

Subtraction. As far as the user is concerned, the operation of subtraction is almost identical to that of addition, the one difference being that a different operation key, marked – or –/= is used. Suppose we wish to subtract *456* from *123*. The number *123* is first entered into the keyboard register then transferred to the answer register by pressing the + key. The number *456* is entered into the keyboard register then transferred to the answer register with the proper sign by pressing = (equals). The answer, *–333* appears immediately. In some calculators the minus sign appears next to the left number; in others it's at the extreme left. The number is called a *"signed number"* to indicate that the sign relates to the number rather than to the operation (subtract).

When working with signed numbers the order of entering the numbers isn't important providing the signs associated with the numbers are entered properly. For example, *123 – 456* may be handled as follows:

enter	press	display
	C	*0*
	+	*0*
123	–	*123*
456	=	*–333*

Alternatively, the operation may be:

enter	press	display
	C	*0*
	–	*0*
456	+	*–456*
123	=	*–333*

Multiplication. Recall that multiplication is just a method of successive addition. Suppose we wish to multiply *456* by *123*. The answer is equal to *456 + 456 + 456*, plus *4560 + 4560* plus *45600*. This can be confirmed by making the addition, the sum being *56088*.

To solve this problem in a calculator, the internal controls cause the multiplicand to be added to itself three times, and the result is stored in the answer register. The multiplicand is then multiplied by *10*, and this value is added to the answer register twice. Finally, the multiplicand is again multiplied by ten, and this value is added to the answer register to get the result, *56088*. This process requires a place to store the multiplicand, a third register, as shown in block form in fig. **2**. In most small calculators the contents of this register can't be displayed so it's often called a *hidden register*.

To the user the operations for multiplication are not much different than for addition. The first number, the *multiplicand* is first entered, followed by x. This operation stores the number in the hidden register. The second number, the *multiplier*, is then entered. Pressing the = or +/= key gives the answer, the *product*.

In very small calculators, it's not possible to multiply

negative numbers directly. The easiest way to handle this is to remember that, from algebra, the product is positive if ***both multiplier and multiplicand have the same sign***; if not, the product is negative.

Overflow. Recall that the number of digits in the product is either equal to the sum of digits of the multiplier and multiplicand, or is one less than the sum. When two large numbers are multiplied, the product may have more digits than the register (and display) capacity. This condition is called *overflow*. It can also occur in addition, since the number of digits in the sum can be one greater than the largest number of digits of the two numbers being added.

Most calculators have some form of indication of overflow. It may be a special symbol at the left of the display, a glowing dot, or a blink. Some small calculators eliminate the need for this by making the answer register twice the size of the keyboard register: pressing a special key marked with a right-pointing arrow causes the additional digits to be displayed. Also overflow can be avoided by a technique called *multiple precision*, described in the next article.

Division: underflow. Just as the calculator handles multiplication by successive addition, it handles division by successive subtraction. For example, to divide *56088* by *123*, the calculator first subtracts *123* from the left 3 digits, *560*, leaving *437*. This is a positive number, so *123* is subtracted again, to get *314*, then again to get *191*, and stil again to get *68*. Another subtraction of *123* causes the sign to change to minus, producing a reading of *-55*, a signal that subtraction has proceeded too far. Accordingly, *123* is added, to get *68* again. This number is shifted one position to the left, bringing in the next digit, for an internal reading of *688*. The net number of subtractions, 4, is recorded as the first digit of the *quotient*. The process of subtract, test, adjust when necessary, then shift continues until the answer is obtained, *456* in this case.

Suppose we wish to divide a very small number by a larger one, say, *3* divided by *987654*. On the small calculators the number appearing as an answer is a zero. This answer is obviously wrong, since the actual answer must be greater than zero. This condition is called *underflow*. It is handled in one of several ways, depending on calculator design. In some, an underflow signal is presented; in others, a special conversion may be used as described below.

Decimals and calculator notation. So far we have ignored the decimal point and have looked at whole number problems. A few types of calculators do the same thing. It's up to the user to keep track of the decimal. This is not difficult if three simple rules are remembered:

1. In addition and subtraction, zeros must be added to the right as needed to make the number of places to the right of the decimal the same for all numbers. Example: 7.6 + 0.25 must be entered as 760 + 25. The indicated answer, 785, is read as 7.85.

2. In multiplication, the number of places to the right of the decimal in the product is equal to the sum of the number of places of the multiplier and multiplicand.

3. In division, a simple method is to use zeros to give the same number of places in divisor and dividend. The answer displayed is the part of the quotient to the right of the decimal point. On most no-decimal calculators, the decimal part of the quotient can then be displayed by pressing a key marked →.

By far the largest number of calculators on the market have decimal provision, indicated by a key labeled with a decimal dot. There are, however, several ways of handling the decimal. Some calculators allow entry of the decimal point at any place up to a limit – often 2, 4, or 5 places. The answer is given to the same number of places (fixed decimal).

Some assume that the number to be entered has two decimal places, as in dollars and cents. Answers show two places (adding machine entry).

Some allow the decimal to occur at any place on the display for both entry and answer. The answer display starts with the first digit, or with a decimal, as necessary (floating decimal).

Some express the answer as a number multiplied by a power of ten, such as 1.2345×10^{-3} (scientific notation).

Some restrict the power of ten to 10^0, 10^3, 10^6 (engineering notation).

In complex machines, there may be manual or automatic change from one notation to the other. The method of handling the decimal is very important, and time should be spent with the instructions and in practice until the mode of operation is thoroughly understood.

simple and extended calculators

So far we've looked at the elements basic to all calculators. As seen with respect to the keyboard, these are:

numericals: *0* to *9*, and . (decimal point)
operations: +, −, x, ÷, =
instructions: C, CE

Depending on the design, these can be placed on 15 to 17 keys, 16 being the most common. In simple calculators only these keys are found. There are however, a number of other instructions and operations that can be added and which will be found in various combinations in the larger and more expensive machines.

Chain calculations. In the simplest calculator the number in the answer register is available only to the display As a result, this register must be cleared at the end of each calculation before starting another to prevent error.

More advanced calculators, by internal switching, make this number available to the keyboard or to the hidden register, to serve as the input for a new series of calculations. This technique is called *chain*, since it allows linking or chaining of successive operations.

Where this method of construction really pays off is in mixed calculations. For example, the problem $\frac{2 \times 3}{4} + 6$ is solved by the successive entries of *2*, x, *3*, ÷, *4*, +/=, *6*, +/=, (calculators with no separate = key), for an answer of *7.5*. Without chain, the entries would be *2*, x, *3*, +/=, C, *6*, ÷, *4*, +/=, C, *1.5*, +, *6*, +/= again for an answer of *7.5* but with the necessity of remembering the result of each step long enough to input it for the next step.

There are some things to be aware of with chain operation. For example if you wish to solve $2^3 = 8$ and you press the keys 2, x, x, x the display will show 16. In the multiply, divide, and add operation above, omission of the intermediate equal sign gives the erroneous result of *6* instead of *7.5*. More important, the order of operations must be correct. In most simple calculators, multiply/divide can be mixed in any order as can add/subtract; but multiply/divide must precede add/subtract in a chain.

Percent. Many small calculators have a key marked with the symbol for percent, %. This is used exactly like the = key to obtain an answer, but it has the effect of shifting the decimal point two places to the right on divide and two places to the left on multiply For example, *2*,÷, *3*, % gives *66.6666*, and *2*, x, *3*, % gives *0.06*. On divide, the result answers the question, "What percent of the second number is the first?" and on multiply, the question, "What is the second number percentage of the first number?"

Values of constants. When a register is cleared, it is actually set to contain zeros. It is no great problem to design it to be set to some other number.

In the small calculator, such "set to a value" capability is rarely provided. Even the larger ones include only a single value, that of π, or *3.141592654*. An exception is the family of special calculators designed for metric conversion. In these, the multiplying factors for feet to meters, pounds to kilograms, etc., are built in and come into play as the appropriately marked key is pressed.

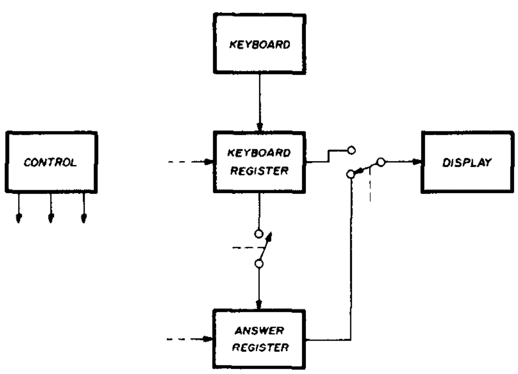


fig. 1. Elements of the basic hand-held electronic calculator. The registers store information input from the keyboard, and the display provides a readout of calculation results based on instructions to the registers from the control circuits.

Change sign. In many small calculators, operations such as division by a negative number are not directly possible, since the – is an instruction to subtract. However, some avoid the problem by providing a special key, usually marked +/–, i.e., *change sign*. This operates only on the keyboard register, causing its sign to change + to – or – to +.

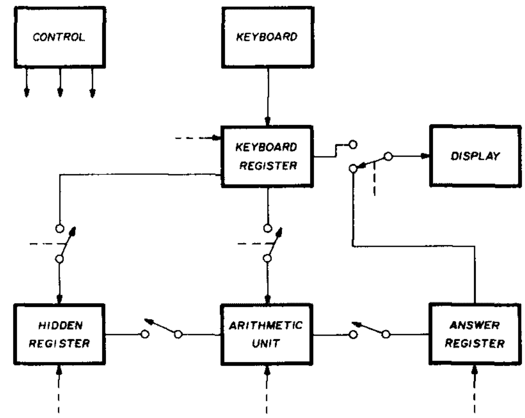


fig. 2. Calculator with add, subtract, divide, and multiply functions. A third register is used to store the multiplicand during problems involving multiplication. Its contents can't be displayed, hence the name "hidden register."

Exchange. In fig. 2, the number in the third, or hidden, register can't be seen; it can't be checked; nor is it available except for specifically intended operations. To increase flexibility, a key may be provided that causes the contents of the keyboard and hidden registers to be exchanged. Usually the key is marked *EX*; i.e., *exchange*. Pressing the key once allows the contents of the hidden register to be examined; a second action returns the contents to the initial condition. Alternatively, after the first actuation, an operation can be performed, say, addition.

Memory 1: constant. One function of the registers shown in figs. 1 and 2 is to remember the numbers placed in them. It seems obvious that calculators would be more flexible in use and would have more calculating power if more memory capability were provided, just as the calculator of fig. 2 has more capability than the simpler unit of fig. 1. In practice this is true.

The simplest type of added memory is called *constant*. When provided, it is usually made evident by a slide switch, one position being marked *K*. When this key is depressed, a number can be loaded into this memory by pressing an arithmetic operation key. Thereafter, entering another number followed by = causes the operation to be performed between two numbers. For example, suppose *constant* is on, and has been loaded by pressing *3.14159*, x. Pressing *120*, =, gives the answer, *376.9908*, or *120π*. Now pressing *2*, = gives *6.28318*. The constant is retained until replaced by keying a number and some operation other than =.

Quite a range of variation is possible in the method of operation. Some calculators require two numbers to be loaded and place the second in the constant memory. In others, the constant memory holds the first entry. Another variation is *automatic constant,* which seems to be coupled with the chain. It operates the same as the switchable constant described above but is always available without switching. In other variations, *constant* can be used with multiply and divide but not with add and subtract. However, despite its exact form, this simple form of memory adds greatly to the calculator power.

Memory 2: four-function memory. A further increase in calculation power can be secured by making the memory more flexible so that it can be loaded or recalled at any point in the calculation cycle. Typically, this is done by providing four keys, *+M, -M, RM, CM,* indicating: add number displayed to the number in memory, subtract number displayed from the number in memory, read memory (transfer number in memory to the keyboard register, leaving the number in memory intact), and clear memory (set the memory number to zero). Some calculators include a flag to show when memory is loaded. Some may omit one or more keys, usually the *-M* key.

The major use of this form of memory is to store intermediate results while making a continuing calculation. For example, to solve $(A \cdot B) - (Y \cdot Z)$, the key strokes are A, x, B, =, +M, Y, x, Z, =, -M, RM. Another use is storage for constants, such as π, or the rate of sales tax, for example.

Memory 3: stacks. As problems become more complex, it is found that several numbers must be "remembered" at the same time. One way of accomplishing this is to provide several memory registers arranged so that each successive call for memory input causes a number already in memory to be displaced into another register. These registers are said to form a *stack.*

Stack memory appears in two forms. One is based on conventional algebraic notation and usually appears on the keyboard as a pair of keys marked [(and)], standing for inner and outer parentheses. Successive keying of the left parenthesis causes the stack shift. The other stack arrangement is based on a special notation called Polish reverse notation (no one can pronounce the mathematician's name!).* The two keys for this are usually marked *ENTER↑*, and *RCL* for recall. A key marked *R↓* for *roll* may be provided to allow reading the memories successively. This type has no equals key; the notation arrangement makes it unnecessary Readout of both of these memories is automatic, occurring as the logic of the problem demands.

Memory 4: addressable memory: programs. As the amount of memory provided is increased, the stack arrangement becomes too limiting. To avoid this problem each memory can be assigned to a symbol, so a number can be placed into any memory cell, or the number in any cell can be read out without calculation. This arrangement is called *addressable memory.*

*The Polish logician, Jan Lukasiewicz.

Addressable memory may be found combined with a special memory, one which remembers the successive steps needed to solve a problem. The assembly of steps is called a *program.* In some calculators the program is stored internally and is lost when the power is shut off. Other calculators store the program on a card, usually similar to magnetic tape. The power of these addressable memory and programmable calculators approaches the capability of computers.

Functions. A function of a quantity bears a definable relationship to the quantity. If X represents the quantity, the quantity $Y = X \cdot X$ or $Y = X^2$ is a function of X.

Small calculators have no built-in provisions for these specific functions. Slightly larger ones may provide X^2, $\sqrt{X}$ and $1/X$, by special keys. Still more powerful ones may provide $\sin X$, $\cos X$, $\sin^{-1} X$, e^x, X^y, etc. These additional functions are generated by special circuits in the control section of the calculator. They are calculated each time they are needed, using the number present in the keyboard register. Usually the calculation takes a noticeable time period, during which the display is blank, or flashing.

The next article in this series describes methods of approximating these functions on the smaller calculators.

which calculator?

Which calculator to buy should be based on expected use, problems to be solved, and cost. A simple calculator is best for simple problems; there is much less chance of error. If more complex problems are to be solved or much repetition is involved, the calculator power should be increased accordingly. However, if complex problems are encountered only at rare intervals, it's best to stay with the simpler calculators, using scratch pads and function tables as substitutes for calculator power (discussed in the next article). Not only is the cost lower, but the time wasted is less – it's easy to forget the tricks of the larger calculators.

The four-function, 16-key calculator with constant is about the simplest usable machine for routine radio calculations. The type with four-function memory (usually 20 keys) is appreciably better. Of course, the regular user should consider the advanced calculator. Regardless of the type used, practice is in order. A good technique for refreshing your memory of problems, and at the same time becoming acquainted with the calculator, is to solve the example problems in *The Radio Amateurs Handbook,* or *The Radio Handbook.* In running these problems, it's a good idea to practice approximate mental solution at the same time – the best way to avoid error.

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ham radio

syllabic vox system

for Drake equipment

Elimination of conventional vox delays is featured in this novel system adapted to the popular Drake T-4XB and R-4B

Radiotelephone conversations without speech delays caused by conventional vox systems is truly a delightful experience. The system described here has been designed for the Drake T-4XB and R-4B transmitter and receiver. It eliminates poor speech habits induced by vox relay delay and also eliminates vox relay contact-bounce transients. In addition, true break-in CW keying is possible with this system.

Dr. Hildreth, WØIP, described his system for instantaneous voice interruption (IVI) in *QST.*[1] While IVI clearly demonstrated the feasibility of rapid-switching syllabic voice response in the control of rf transmission, I experienced difficulties when trying to use conventional vox with IVI.

conventional vox systems

Vox, anti-vox, and delay circuits in most popular transmitters are similar in design principles. Therefore, these techniques might be called "conventional vox." Conventional vox depends on passive integrals developed from the audio signal by rectified dc charges to a control capacitor. Output from the capacitor appears at the grid of a relay-switching tube; or, in solid-state circuits, as current to the base of a relay switching transistor. When a positive charge provides vox activation during transmission, anti-vox during reception requires a negative charge to inhibit operation of the vox relay.

A significant period of time is required for both vox and anti-vox functions because the control capacitor can't be charged or discharged instantaneously. Time, dependent on source and load resistance, together with the value of the control capacitor, is extended accordingly. Likewise, time is equally important to the anti-vox input, which requires similar processing of opposite polarity. A delay of many milliseconds, or even seconds, may be involved, which varies with the amplitudes of vox input and anti-vox receiver output. Usually, these functions are mutually dependent. Such factors contribute to annoying and compromised operating adjustments, often leading to abandonment of vox in favor of PTT.

In the use of conventional vox systems, the operating-point bias to the final amplifier tubes is active throughout the vox relay *on* time. This, of course, means that final-amplifier bias current is present between speech syllables, words, and code elements. Power used in this manner, without rf transmission, is wasted. Also, a T-R switch loses merit, because final-amplifier tube noise appears at the switch and receiver input.

Both these undesirable conditions are eliminated by the syllabic vox method. Final-amplifier operational bias is applied only during rf transmission. At all other times, the final amplifier tubes are cut off completely and no plate current flows. Significant improvement in transmission efficiency results.

adaptation to Drake equipment

Using this system with the Drake R-4B and T-4XB requires no hole drilling or component changes. The only outboard unit required is a small circuit board containing the 5-volt power supply, which is mounted in the Drake MS-4 speaker cabinet. During operation, all controls and transmission modes are in accordance with the Drake instruction manual.

All components except the T-R switch and power supply are mounted on a 3.5 by 4.75-inch (89 by 121 mm) Vector board. The components are self supporting on the board and no heatsinks are required. The circuit board is mounted in the upper right-hand corner of the T-4XB transmitter, where adequate space is available without crowding. Two small L-brackets secure the board to the chassis. Existing chassis screws secure the board. No critical or difficult wiring problems were encountered. Total component cost is about $50.00.

By R.W. Hitchcock, W6RM, 667 West Arrow Highway, Upland, California 91786

Referring to **fig. 1**, all logic components are in the 7400-series ICs, U2, U3, and U4. The retriggerable monostable, U3, is the most important single component in this circuit. With the RC values shown, a single positive pulse at U3A pin 3 would deliver a *low* output at U3A pin 6 for 8.4 milliseconds, plus or minus the tolerance of the RC combination. However, the device offers retriggering capability at all audio frequencies between 120 and 4000 Hz in this application, which means that pin 6 of U3A or U3C will remain low during voice syllables. Upon cessation of voice signals, U3A-6 or U3C-6 will go high after 8.4 milliseconds, which is the time required for one CW dot at 140 wpm. Obviously, this cutoff time is insignificant with respect to voice syllable duration.

U1, an LM3900 quad operation amplifier, was chosen for front-end application to both vox and anti-vox channels. In each case, the first stage offers negligible loading to a high-impedance source and delivers approximately unity gain throughout the range of the source. The second stage produces the required signal gain and stabilization. Use of this amplifier group assures circuit reproducibility with no concern for the varying parameters of discrete devices.

Output from each two-stage amplifier is introduced directly to Schmitt triggers U2A and U2B for square-wave shaping necessary to following logic. The Schmitt triggers have hysteresis values of nearly one volt between upper and lower trigger points. This characteristic is quite important to noise immunity when setting vox and anti-vox gain during operation. Signal inversion occurs in the vox amplifier channel, but no inversion is present in the anti-vox amplifier.

These conditions are essential to the positive CW start pulse and the negative CW stop pulse but they have no influence on radiotelephone transmission. In this way, activation and inhibition of CW vox occurs at the start and stop points of all code elements. Thus true break-in follows with only the inherent delay of the receiver mute circuit.

One section of the 7400 NAND gate, U4, provides a low at either U4A pin 1 or U4A pin 2, which will deliver a high at U4A pin 3. This condition is essential to the solid transmission *on* requirement for *tune,* and *PTT,* while retaining the syllabic voice response capability of normal vox transmission.

U4A drives U4B, an inverter, which activates Q1 into the open-collector cutoff condition during transmission and into saturation during reception. Under the open-collector transmission mode, 50 mA flows through CR5 and the 2.5 mH rf choke in the T-R switch. This action places the node at the junction of C11, CR5, C12, and the rfc at virtual ground during transmission. In the receiving mode, this nodal point offers no loading to the received signal other than the desired receiver input. Diodes CR4, CR6, and CR7 are protective devices in the event forward bias fails.

Output of CR9 from Q1 also drives the base input of Q4 to saturation during transmission. This condition allows Q5 to operate in the open-collector mode and permits full muting to the receiver through the mute line. While receiving, Q4 is cut off. Now, current flows through the 430-ohm collector resistor and CR15 to saturate Q5. The mute line is brought to ground in this manner for normal reception by the R-4B.

Transistor Q4 also drives Q6, which is emitter-biased to +1.5 volts by CR17 and CR18. When Q4 is saturated during transmission, Q6's base goes negative with respect to the emitter and saturates. When saturated, the collector potential of Q6 is approximately +1.5 V with respect to the collector of Q7. This action results in saturation of Q7, which removes cutoff bias to a kW power amplifier, if one is used.

Returning to the output, U4A pin 3, this point also triggers U3B pin 3. The output at U3B pin 6 now goes low for a constant period of 100 microseconds. With U3B pin 6 low, U4D pin 13 is held low and U4D pin 11 remains high during the 100-microsecond delay. The high at U4D pin 11 holds Q2 at saturation with a collector potential approaching ground. This condition holds cathode switch Q8 at cutoff for the delay period. The 100-microsecond delay for the T-4XB final amplifier is not significant, operationally. However, it provides full assurance that the T-4XB rf output cannot appear at the T-R switch until 100 microseconds *after* the T-R switch, mute line, and kW final-amplifier bias have been activated for transmission.

Both U4D pin 12 and U4D pin 13 must be high to force a low at U4D pin 11 during transmission. During reception U4B pin 6, driving inverter U4C, delivers a low at U4C pin 8. As long as U4C pin 8 remains low, no change of state at U4D pin 13 can alter the high at U4D pin 11. However, when both U4D pin 12 and U4D pin 13 are high, a low will hold at U4D pin 11. Propagation time for a change of state through U3B is somewhat greater than that of U4B and U4C. So, a small delay in the change from low to high at U4D pin 12 is desired. This delay is accomplished by means of a 270-ohm resistor and C15 (0.01 μF), which are shown at the output, U4C pin 8. This step permits propagation time for U3B and eliminates the possibility of a short, unwanted low transient at U4D pin 11.

The anti-vox signal is picked up through an extension cable from a point near the anti-vox gain pot in the T-4XB, and routed to the circuit board. This signal performs the inhibit function to U3A. If the output at U3C pin 6 goes low, any low existing at U3A pin 6 will promptly go high, having been inhibited by U3C. Also, when U3C pin 6 is held low, no activation of U3A can occur. This is an important and useful feature of the retriggerable monostable. Transmission, then, must be initiated between syllables and words. This is a normal occurrence in landline telephone conversations. The anti-vox circuit is quite similar to the audio and sidetone circuit, making further description of anti-vox unnecessary.

Note that no variable adjustments by pots and no component tailoring are required in this design. However, the normal T-4XB vox gain and anti-vox gain controls are used for noncritical settings of each function. The vox delay pot in the T-4XB is not used. In operation of the new vox method, no time delays are

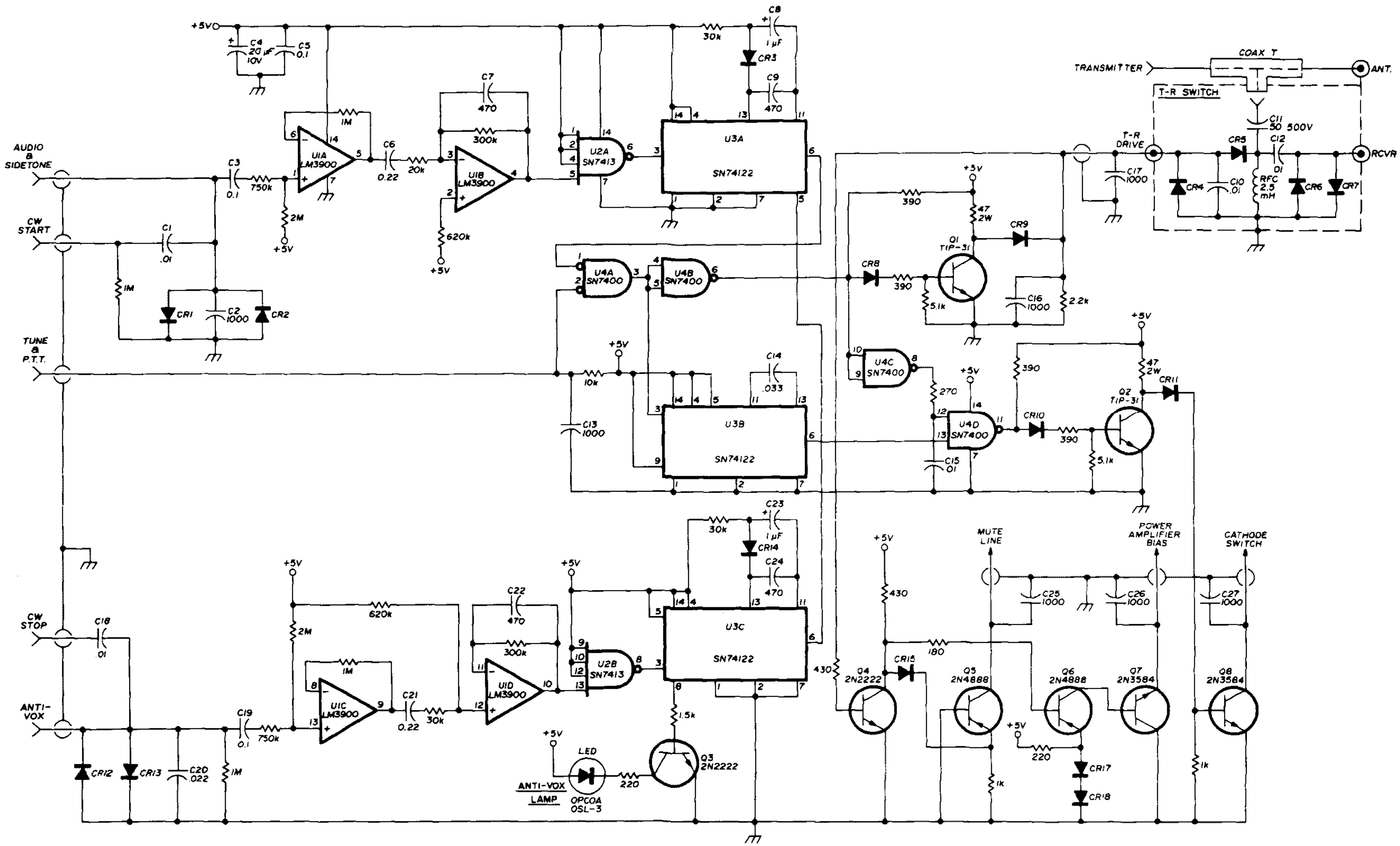

fig. 1. Syllabic vox and CW break-in system schematic for use with the Drake T-4XB and R-4B transmitter and receiver. Circuit provides instantaneous speech communication without conventional vox relays. Diodes CR4, CR5, CR6, CR7, CR9 and CR11 are 1N4004 or equivalent; all other diodes are 1N914 or equivalent. The rf choke is a 2.5 mH transmitting type (100 mA). All resistors are ½ watt, 10% unless otherwise specified.

present that can be observed by unaided human senses. Also no vox/anti-vox interaction occurs. During the initial setup, the threshold for vox activation is established by the vox gain adjustment. After this setting, an increase in voice level does not alter the vox response in any way. Similarly, the anti-vox signal threshold assumes a constant level for activation after adjustment of anti-vox gain.

setup procedure

The setup procedure is no more difficult and is perhaps simpler than that of conventional vox.

Step 1. With the receiver and transmitter in the ssb mode, the anti-vox gain, vox gain, and receiver audio gain pots should be set to minimum gain. With the voice at normal audio level and the microphone in its customary position, bring up the vox gain sensitivity only enough to observe vox enabling, which will be shown by a fluttering action of the S-meter between syllables and words. A sustained audio note will hold the S-meter at maximum indication. Do not change this setting of the vox gain pot in the next step.

Step 2. With receiver band noise only, turn up the receiver audio gain until the vox is triggered occasionally, which is indicated on the S-meter as in the first step. This audio gain level should be high relative to that used in normal signal reception. Continuing with the band noise conditions, increase the anti-vox gain until the anti-vox LED glows intermittently, while the S-meter remains at zero.

The anti-vox LED is useful when communicating with a very weak signal under high-level noise conditions. These conditions combine to present a "worst-case" operational example. In this situation, it is advisable to reduce receiver audio gain until the anti-vox LED does not glow intermittently due to adjacent frequency audio interference.

CW transmission

The CW start pulse in the T-4XB is used for fast vox initiation upon key or keyer contact closure at the start of a dot or dash code element. After starting, the sidetone signal holds the vox system *on* throughout the time duration of a dot or dash. However, the end of a dot or dash must inhibit the vox circuit promptly to prevent holding the mute line open for 8.4 milliseconds after the end of a dot or dash. This requirement is accomplished by negative pulse differentiation upon restoration of the grid bias line in the T-4XB. Then, the negative-going CW stop pulse is routed to the circuit board anti-vox circuit, which inhibits U3A.

power supply

The T-4XB normally uses a 6EV7 for vox gain and vox relay driving. Neither the vox relay nor the 6EV7 are used with the new vox system. So a simple heater wiring change is made that eliminates the 6EV7 heater load of nearly 4 watts. The new circuit board requires a separate, regulated power supply that delivers 5 volts dc at 425 mA. Elimination of the 6EV7 suggested use of the Drake AC-4 power supply for the 5 Vdc requirement without adding an additional load to the AC-4 unit. Consequently, a 5-Vdc power supply was designed for use with the 12.6 Vac transformer winding of the AC-4 (**fig. 2**).

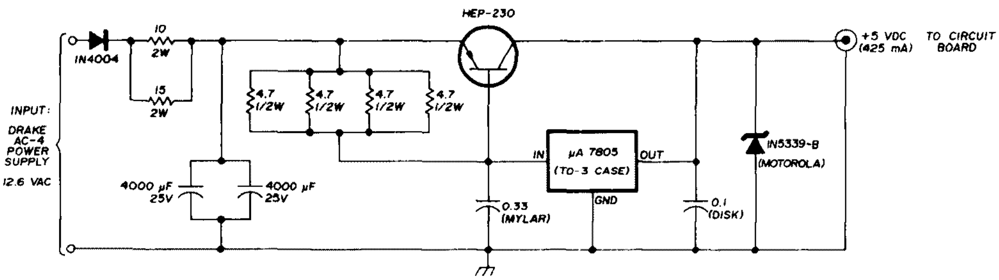

fig. 2. Simple 5-volt power supply for the syllabic system. Input power is supplied by the Drake AC-4 power supply. Unit is installed in the Drake MS-4 speaker cabinet.

A small circuit board was prepared to support all components of the 5-Vdc power supply. Then 12.6 Vac and ground leads were soldered directly to these transformer terminals inside the AC-4 unit. The new leads were brought out of the AC-4 case through ventilation holes and connected to the input of the 5-Vdc power supply This new supply was then located between the AC-4 power supply and the speaker in the Drake MS-4 speaker cabinet. This location offers more than adequate space for the 5-Vdc power supply, and there is no temperature problem. Incidentally, this method of connecting the available 12.6 Vac power automatically places the 5 Vdc power under the control of the power *on/off* switch on the T-4XB panel.

T-R switch

The T-R switch used in this system is a compromise with respect to electromechanical relays. The titles "T-R switch" or "diode switch" are erroneous in technical fact. But these titles are used since correct and definitive nomenclature is not immediately available. Casual observation of the circuit may lead to the false conclusion that the T-R switch is a simple device for receiver protection. However, a rigorous, quantitative analysis of this

device is beyond the scope of this article. The interested reader is invited to bypass the T-R switch with a continuous 50-ohm line from antenna to receiver for comparison testing. Such tests conducted at my station displayed no discernible difference of signal strength on the S-meter.

Like other electronic T-R switches, this switch must be handled properly to avoid signal suckout. In this case, the switch enclosure is mounted directly to the T-4XB output connector. Rigid coaxial hardware is used and provides secure mechanical support for the small T-R switch can. No signal suckout will occur with this installation of the switch unless the receiver and transmitter are tuned in different bands. When operating normally in a given band, improvement in signal-to-noise ratio may be observed. This is because the received signal looks into the output tank of the transmitter before appearing at the receiver.

T-R switch with linear amplifier

When a final amplifier is used, the dual T-R switch method should be used, and the second T-R switch should be mounted rigidly to the output of the kW amplifier. See **fig. 3** for a schematic and component details. In this case, two changes must be made in the control of the kW amplifier:

First, the power amplifier bias control must be removed from the antenna relay section that normally switches cutoff bias control to ground. Then, the bias control lead should be a direct, isolated conductor from the bias control point in the linear amplifier to the power amplifier bias terminal on the new circuit board in the T-4XB. It is assumed that the bias control point in the final amplifier has a negative potential not exceeding −150 Vdc and current loading not greater than 150 mA when grounded. Positive cathode bias is not considered in this article.

Second, the antenna relay in the kW amplifier must be wired so that *on/off* control of this relay is performed by the *power on* switch on the kW amplifier panel *only.*

These changes were made to an NCL-2000 amplifier in less than one hour. The end result was a pleasing lack of clacking noise from the cumbersome relay in dynamic operation.

T-4XB changes

The following steps define installation of all extension leads, minor changes, and cabling required to the T-4XB chassis. No holes are drilled, no switches added and no defacement of components or chassis are required. (The Drake T-4XB manual is referenced).

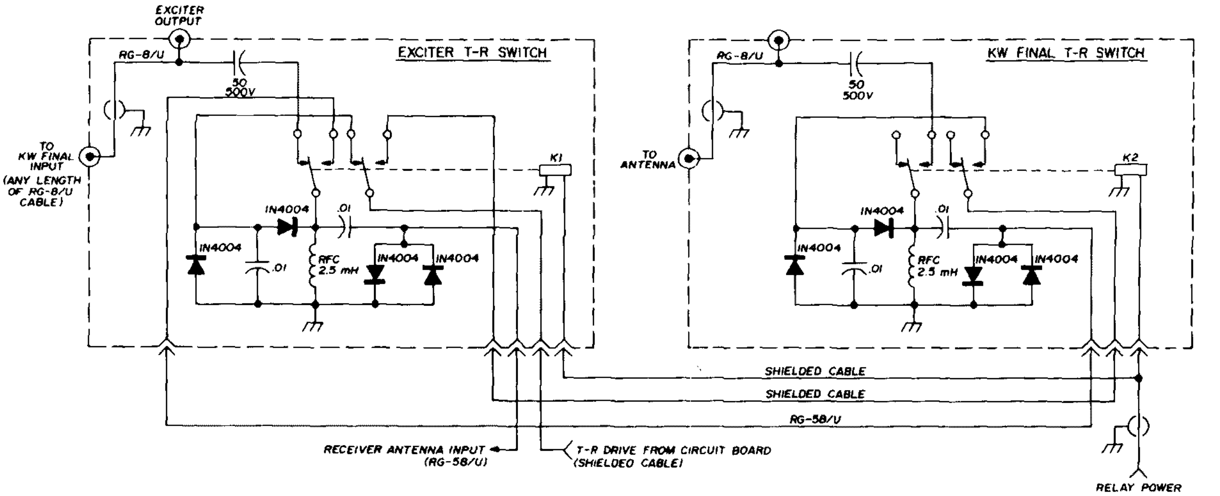

fig. 3. T-R switch schematic for use with kilowatt final amplifiers. Suggested relays are Potter & Brumfield type KT-11D. Each T-R enclosure should be mounted to each output by rigid coaxial connectors. A bud type CU-234 utility box is suggested for each T-R switch.

1. Connect a jumper wire between socket terminals 4 and 5 of V10.

2. Disconnect lead from pin 5 of V9 to pin 3 of V11.

3. Disconnect lead from pin 9 of V9 to ground.

4. Connect new lead from pin 9 of V9 to pin 3 of V11.

5. At the vox relay *ant* section to J7, connect a jumper wire between the normally open contact and the movable contact. Then disconnect the lead to *rcvr* J6 from the normally closed contact.

6. At the vox relay mute section to J8, disconnect the movable contact from ground. Connect a board extension lead labelled "mute line" to the normally closed contact. Then connect another extension lead labelled "cathode switch" to the normally open contact using no. 22 AWG (0.6 mm) or larger shielded wire.

7. The third section of the vox relay is not specified but is used to control the grounding of kW-amplifier negative bias in this application. Connect an extension lead labelled "power amplifier bias" to the normally open contact of this section. The 2N3584 will easily handle a 150-mA load at −150 Vdc if required.

8. Locate the PTT contact on microphone jack J3 and connect an extension lead labelled "tune and PTT" to this contact.

9. Locate diode CR7 on the small circuit board at the right rear section of the T-4XB chassis. Apply soldering iron and pull out the cathode end of CR7. Then connect an extension lead labelled "CW start" to the exposed cathode of CR7.

10. The anode of CR7 connects to capacitor C131 (0.01 μF), and the other side of C131 is connected to R78 (1M). At the junction of R78 and C131, connect an extension lead labelled "CW stop." Now solder a 1000-pF disc capacitor across R78. (Incidentally, this is the only component added to the T-4XB chassis).

11. Locate C92 (0.02 μF) and CR12 in the anti-vox circuit. Then connect an extension lead labelled "anti-vox" at the junction between C92 and CR12.

12. C142 (0.02 μF) is connected at the junction of R98 (3.3M) and CR9. Connect a jumper across C142.

13. Connect an extension lead labelled "audio and side-tone" to the center wiper contact of the vos gain pot, R89.

14. A shielded cable is connected to the T-R terminal on the circuit board and routed outside the T-4XB cabinet to the designated terminal of the T-R switch. The cable delivering +5 Vdc power to the circuit board and the cable to the T-R switch may be held together and clamped by a convenient screw already located on the T-4XB chassis.

This completes all changes and extension-lead connections to the T-4XB chassis. None of these terminals lack adequate access and offer no problem in location. At trade-in time, the T-4XB may be quickly restored to its original design with no loss in value.

results

In radiotelephone practice, you'll experience some new and pleasant features when using this system. There is no vox-on presence between words or after a word of transmission. Only words and syllables of words appear on the air. Also there are no vox rf transients. The operator may be broken at any time and "doubling" is eliminated. Extraneous local noise does not appear on the air because the voice frequencies overwhelm the noise. If the transmitting operator wishes to force transmission, he may use PTT in the normal manner.

In CW service, the only locally generated noise is from the handkey, keyer, or power-amplifier blower. Only sidetone is heard in the earphones or speaker. Clicks and TVI are absent.

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ham radio

electrical units:

their derivation and history

An interesting sidelight on measurement units and their relationship to the metric system

A lot of people who work in electronics have to deal on a day-to-day basis with quantities such as voltage, resistance, and current and the units used to measure these quantities. Surprisingly few, however, know how the units of measurement were established, and it almost never fails to raise an eyebrow when they find out that all the basic quantities – volts, amperes, etc – are metric units that can be defined in terms of kilograms, seconds, and meters.[1]

It's strange to think that the amount of inductance we call 1 henry can be traced back to the diameter of the earth, or that 1 farad of capacitance owes part of its definition to the weight of water. It's stranger yet to realize that these quantities wouldn't have the values they do if it hadn't been for the downfall of King Louis XVI in the French revolution of 1789 – but that's how it is.

basic units

The watt, as a unit of power, and the ampere, as a unit of current, provide the foundation for defining the other units in electronics. Watts and amperes owe their definitions to the field of physics, which we'll look at later. The ohm, as a unit of resistance, arises naturally from the ampere and the watt. It was found that only a specific amount of resistance would produce 1 watt of heat if 1 ampere were flowing through it. This value of resistance is *defined* as 1 ohm, and the potential drop developed across this resistance under these conditions is *defined* as 1 volt. Even though an infinite number of combinations of voltage and current will produce 1 watt of heat in a resistor, there is only one value of emf that will do it with 1 ampere of current, just as there is only one value of resistance that will draw 1 ampere with 1 volt across it. Accordingly, there is no ambiguity in the values: they are clearly defined and have no "double values."

The ampere, as a unit of current, is commonly thought of as a quantity, but actually it is a *rate*, or quantity per unit of time. Its hydraulic equivalent would be gallons per unit of time, or liters per unit of time if you want to stay in the metric system. The electrical equivalent of these gallons or liters is the coulomb, which signifies the actual quantity or number of electrons that will sustain 1 ampere of current for 1 second. In round numbers, 1 coulomb equals 6.25×10^{18} electrons. It's more commonly called the unit of charge, and since it defines a quantity instead of a rate, its symbol is Q. Incidentally, the ampere was originally described as the unit defining the *intensity* of a current, hence the use of the letter I in Ohm's and Kirchoff's laws to symbolize it.

derived units

The units of capacitance and inductance owe their definitions to the volt, ampere, and second. In the case of capacitance, it was discovered that no current would flow into or out of a capacitor unless the voltage across it was *changing.* A further investigation revealed that if the voltage changed at a constant rate, the current would also be constant. If the voltage increased, the current would be a charging current; if it decreased, the capacitor would discharge current back into the circuit. A still closer look revealed that, by doubling or tripling the rate of voltage change, the current would proportionately be doubled or tripled. As a result of these discoveries, 1 farad of capacitance is *defined* as the amount of capacitance that will draw a 1-ampere charging current if the voltage across it increases at the rate of 1 volt per second.

Inductance was found to have similar qualities, but with the roles of voltage and current interchanged. That is, no voltage would exist across a coil unless the current flowing through it was changing. If the current was increased at a constant rate, the voltage developed across the terminals of the coil would also be constant; and doubling or tripling the rate of current change would proportionately double or triple the voltage developed across the coil.

One henry of inductance was therefore *defined* as the amount of inductance that would sustain a 1-volt drop across its terminals if the current increased at a rate of 1 ampere per second.

Thanks to this system of definitions, proportionality constants are conspicuously missing in the equations used in electronics. The factor 2π, which crops up in the equations for reactance and resonant frequency, is the only exception; and even here it is unnecessary if the frequency is expressed in terms of radians per second instead of cycles per second – but that's another story.

power and energy

It was stated earlier that the ampere is not a quantity but is a rate of change per unit of time. It would be just as correct to say that a given current was 50 millicoulombs per second as it would be to call it 50 milliamperes. The watt, as a measure of power, is also a measure

By Robert R. Simmons, WB6EYV, 1640 Walnut Avenue, Long Beach, California 90873

of rate instead of quantity. Just as 1 coulomb per second is commonly called 1 ampere, 1 watt could also be called 1 joule per second. Here arises the distinction between power and energy: energy is a *quantity*, whereas power expresses the *rate* at which this energy is expended.

An example would be appropriate here. Suppose we have a quantity of water at room temperature and wish to heat it to 10 degrees above room temperature. Let's further specify that the quantity of water is chosen so that it will require 5000 joules of heat energy to do this. We could put a resistor into the water and adjust the power fed to the resistor so that it dissipated 50 watts of heat into the water. At this rate, it would take 100 seconds to heat the water. If the power were increased to 200 watts, it would take only 25 seconds; at 500 watts, 10 seconds; and so on. In each case the same amount of work is accomplished; the only difference is in the rate at which it was done and the amount of time required. The joule is a unit of work or energy and has its definition in the roots of the metric system.

measurement systems

A few words about measurement systems are in order. Basically, there are three major systems used in the world to measure quantities and rates on a scientific basis. These are a) the familiar English system, based on the foot for length, the pound as a unit of *force* (not mass), and the second for time; b) the CGS system; and c) the MKS system.

The CGS system (centimeter-gram-second) is the system that Einstein used in his work in relativity. The MKS system (meter-kilogram-second) is the system on which electronic units are based. These last two systems are metric and are based on the meter, gram, and second, which were established in France as a new system of measurement after the downfall of the ruling nobility in 1789. The meter, as a measure of distance, was defined at the time as representing one ten-millionth of the distance from the North Pole to the equator. The gram was also defined as the amount of mass represented by 1 cubic centimeter of distilled water cooled to 4 degrees Celsius. (This is the temperature at which water is most dense). As a unit of time, the second was retained, having its definition in the motion of the earth. The liter, as a unit of volume, was incidentally defined as representing 1000 cubic centimeters.

force and work

In the MKS system, the first derived unit of measure is the newton, which is a measure of force, and is defined as follows: If a 1-kilogram mass were placed on a frictionless surface, this mass could be accelerated from a dead stop to any speed by exerting force on it. The greater the force, the more quickly the mass would accelerate. One newton of force, if exerted on this 1-kilogram mass for 1 second of time, would bring it from a dead stop to a speed of 1 meter per second.

The unit of work in the MKS system is the joule and derives its meaning from the newton and the meter. It is defined as the amount of work expended by moving an object across a rough surface for a distance of 1 meter, if 1 newton of force is required to move it. If the object requires 2 newtons to move it, and it is pushed 2 meters, the work would be 4 joules. The work expended equals the product of the force and the distance.

Ham Radio is designed to provide something for everyone. This broadbush treatment of electrical units, their derivation, and history is presented to encourage further reading in the world of physics, the basis of our electronic heritage. A more rigorous definition of international units including a table of physical constants with their symbols, values, units, prefixes and least-squares error adjustments appears in reference 1. Also included in reference 1 are ten pages of conversion factors to get you out of the awkward English system and into the scientific metric system of measurement. **Editor**

power

Notice that the amount of time required to perform this work does not change the amount of work that is done. Whether the work is done slowly or quickly, the same amount of energy is expended to move the object. The rate at which this work is done is called the *power*, or energy per unit of time. In the first example above, 1 watt of power would accomplish the job in 1 second. In fact, this is the definition of 1 watt: expending energy at the rate of 1 joule per second. In the second example, 1 watt would accomplish the job in 4 seconds, 2 watts in 2 seconds, or 4 watts in 1 second.

So much for the newton, joule, and watt as mechanical units. Only the ampere now has to be defined in mechanical units to completely relate all the electronic units to the world of physics. The ampere is defined under these conditions: If two very long wires (theoretically, infinitely long) are placed parallel to each other and spaced 1 meter apart, and if current is run through them (same value of current for each wire), then the magnetic fields they set up will tend to force the two wires either together or apart, depending on whether the currents are flowing in opposite or in the same direction, respectively.

If the amount of this force exerted on a 1-meter length of either of the wires has a magnitude of 2×10^{-7} newtons then the current is defined as having an intensity of 1 ampere. I don't know why 1×10^{-7} newtons wasn't chosen to define the force created by one ampere, but perhaps it was thought that each wire actually does exert this force, so that the total force acting on a 1-meter length would be 2×10^{-7} newtons — half of it due to each wire's magnetic field. This is all pure speculation, however, and should not be construed as accurate.

I hope this article has shed some light on our heritage in the field of physics without too much confusion. In a country that must go through the withdrawal pains of the English system of measurement, it's reassuring to know that we work in a field that has always been metric.

reference

1. *The International System of Units,* NASA Publication SP-7012, Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402, price 30 cents.

ham radio

i-f/detector receiver module

Design and construction data for an i-f strip, detector, and audio filter that can be used for a high-performance amateur-band receiver

This article provides design and construction details for a high-performance i-f/detector *receiver module*. By *itself* the circuit is an 80-meter-band receiver, requiring only a frequency-determining oscillator and an audio stage for on-the-air use on the high-frequency amateur band. Performance data is given in table 1.

The sensitivity of this receiver is less than 0.1 microvolt rms for a signal-to-noise ratio of 10 dB. What does this really mean? Looking through the ads, you'll find a number of claims as to the sensitivity of a receiver at some signal-to-noise ratio. What's most important (and always missing in these claims) is a definition of terms. For example, one manufacturer tells us that his receiver has a sensitivity of 1 microvolt for a 10-dB signal-to-noise ratio. The question is, "Is this 1 microvolt rms or 1 microvolt peak-to-peak?" By proper attention to details in the construction of the receiver described here, you should be able to measure its sensitivity at less than 0.1 microvolt *rms* for a 10-dB s/n ratio, not 0.1 microvolt peak. At 1 microvolt rms, you should be able to measure a 10:1 voltage difference (20 dB) between signal and noise. I made my measurements using an HP-606A signal generator and a Tektronix 513 oscilloscope.

design features

The block diagram of the i-f/detector (fig. 1) illustrates the overall design. Good image rejection is obtained by initial amplification at the higher frequency followed by a mixer stage with low noise and modest conversion gain. Amplification and passband characteristics are obtained with a multipurpose IC, using either a crystal or ceramic filter, or an LC tuned circuit at the conversion frequency. In any receiver system the detected audio will contain beat oscillator and spurious noise components; these are reduced by passing the signal through an active low-pass filter IC, which is matched to an IC amplifier stage.

The i-f/detector schematic is shown in **figs. 2A** and 2B. Values for resonant-circuit components are shown in table 2. Note that in the schematics, capacitors are shown as a fraction; for example, 0.1/C. This means that the capacitor is a 0.1 μF ceramic. An M below the value means the capacitor is mica.

The 3.5 to 4.0 MHz rf signal is amplified by Q1 and Q2. Tuning of both stages is accomplished using a varactor diode, which eliminates the need for a large variable capacitor. Dual-gate mosfets allow external agc and rf gain control if desired. For applications where an agc and rf gain control isn't required, a single-gate device could be directly substituted, or the control gates, G2, of Q1 and Q2 may be tied to some fixed voltage value. Down-frequency conversion is accomplished by Q3, a standard mosfet mixer. The input tuned circuit of Q3 is fixed and broadly tuned; the output is matched to a commercial i-f transformer. Biasing of gate 2 is accomplished by using a gate-to-source resistance that establishes the operating point for the device.

Overall system voltage gain, detection, and passband characteristics are developed by U1, a multipurpose IC. The schematic illustrates a Collins mechanical filter for

table 1. Performance data for the high-performance i-f/detector module.

frequency range	3.1 to 4.5 MHz with suitable hfo
nominal operating frequency range	3.5 to 4.0 MHz
high frequency oscillator range	3.045 to 3.545 MHz ≈ 2V, peak-to-peak
sensitivity	<0.1 μV rms for 10 dB s/n ratio (distinguishable signal of 1 kHz)
noise margin	>20 dB at 1 μV rms
shape factor	dependent upon filter, Normally 60 dB with Collins of Murata ceramic filters
in-band spurious signals	none using chassis and shields as indicated
audio	0 to 3 kHz; 5 kHz notch; 15 to 20 dB minimum attenuation on signals >5 kHz
nominal gain	80 to 100 dB depending on link positioning and *control voltages*

By M.A. Chapman, K6SDX, 935 Elmview Drive, Encinitas, California 92024

passband shaping. You may prefer a ceramic or simple transformer-type component instead; the overall operation of U1 will remain the same regardless of the type of filter used. Variations in gain can be expected using crystal and transformer filters, which have operating impedances differing from the ceramic type shown.

output of the stages arranged along the LMB 850 chassis sides through feed through terminals.* This circuit is a classic i-f amplifier approach that will provide excellent interstage shielding and minimum feedback coupling which will eliminate oscillation. Interstage shields are aluminum strips fastened to the PC board and chassis

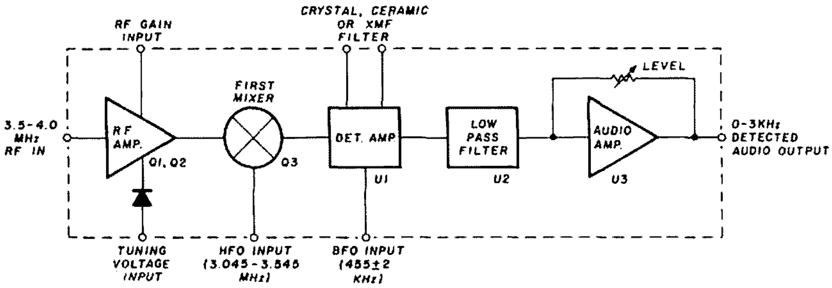


fig. 1. Block diagram, high-performance i-f/detector receiver module.

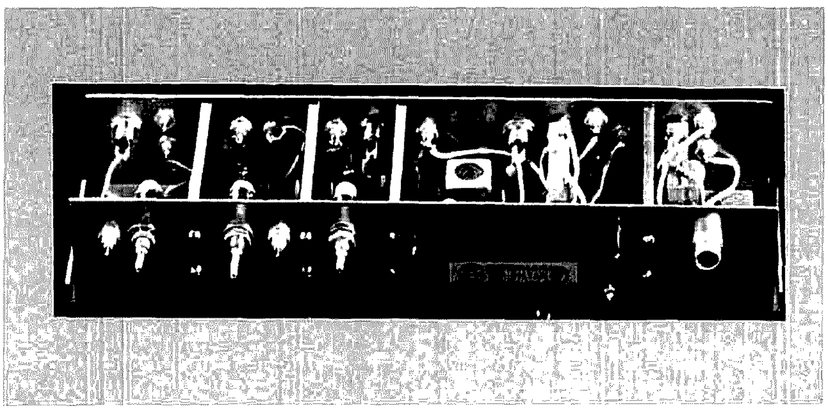
The detected audio signal on U1, pin 7, will contain many beat-oscillator audio components or hiss-type noises in addition to the desired detected audio. The most common method of eliminating these undesirable noise products is to use a toroidal LC or multiple op-amp active low-pass filter. These methods work well but require large areas on the chassis and can be expensive.

An inexpensive active device is available that operates as an elliptic lowpass filter having a rolloff frequency of 2.5 kHz. The attenuation at 5 kHz is approximately 18

sides. These shields provide all the mechanical rigidity required.

There are two methods of building the i-f/detector module: the right way and the sure-failure way. The right way is to build the assembly one stage at a time, testing each stage as you go. The sure-failure way is to install all components, assemble, and apply power expecting it to work!

Construction should start with U3. Install its associated components, temporarily apply power and signal



High-performance i-f/detector receiver module, cover removed. Enclosure is an LMB 850 box. Rf input is at right; audio output at left.

dB with increasing attenuation of all higher-order components. U2's output is matched to the input of U3, which operates as a voltage amplifier to compensate for the loss in gain through U2. U3 is a straightforward operational amplifier with internal frequency compensation.

construction

The photographs show the construction approach. A PC board carries all active components with input-

*Printed-circuit boards are available from the author for $3.00 plus postage.

inputs, then verify the correct output signal. The next step is to install U2, its associated components, and temporarily apply power and signal levels to U2 to verify its operation. This procedure should continue with U1, Q3, Q2, and finally Q1. Proceed through each stage to ensure proper operation. Using the step-by-step approach, you'll become familiar with circuit operation and can optimize the stages for your own preference.

Figs. 3A and 3B illustrate the PC board component locations. The board is designed to accomodate ¼-watt resistors and low-voltage capacitors; larger components may be used by mounting them vertically with respect to the PC board.

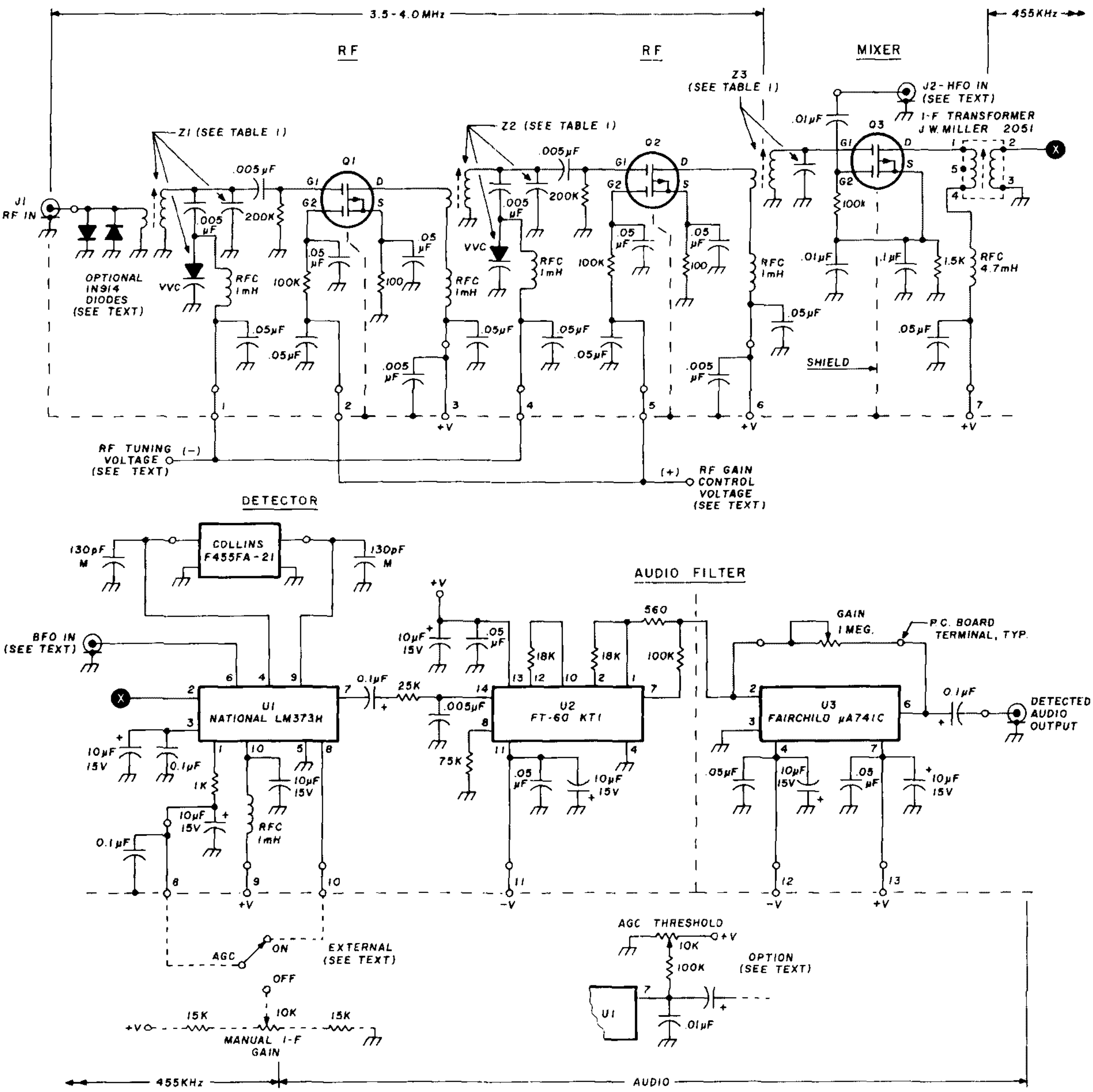
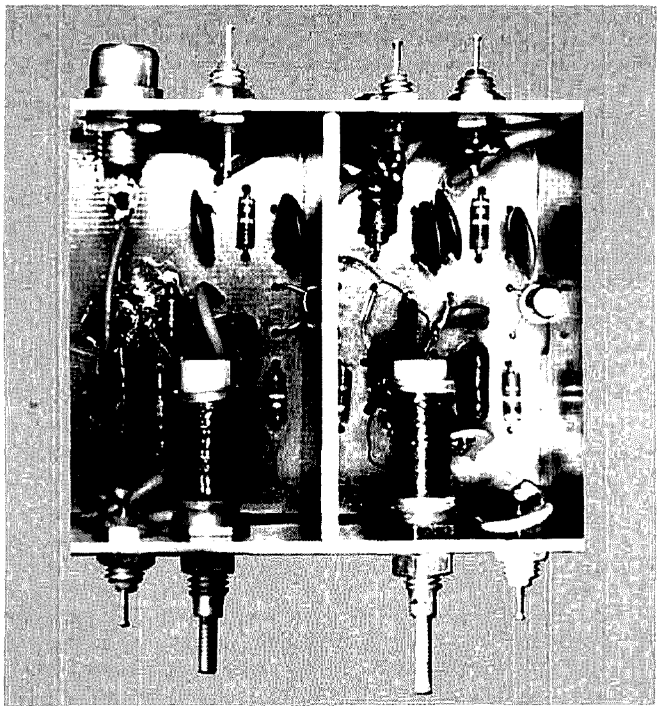


fig. 2. Schematic diagram of the high-performance i-f/detector module. Tuned circuit values are listed in table 2.

step 1. U3 installation. Install U3 and all associated fixed PC-board components. The 1-megohm feedback resistor may be left freely supported over the top of U3 using bus wires from the PC board. Temporarily install one end of a 100-ohm resistor in the +12 and −12V terminals. These resistors limit the current during test in the event of a wiring or circuit error. Apply a 15mV 1-kHz audio signal to pin 2 of U3, and with power applied, adjust the feedback pot so that a 15-volt peak-to-peak signal appears at pin 6; this signal may be taken from the output side of 0.1/C capacitor of pin 6.

step 2. U2 installation. Install U2 and all associated components up to the 0.1/C coupling capacitor from pin 7 of U1. Install the 100-ohm current-limiting resistors for both the +12 and −12V terminals, as you did for U3. With power on both U2 and U3, apply a 100mV 1-kHz audio signal to the U1 pin 7 side of the 0.1/C capacitor. While monitoring the output of U3, adjust the U3 feedback pot for an output voltage of 100 to 150 mV. Increasing the frequency of the audio input signal should show decreasing output values at frequencies above 3 kHz.

U2 can be optimized by adjusting the resistor from pin 8 to ground and the 560-ohm series output resistor from pin 1, so that gain and rolloff slope are sharpened. The CW buff will probably want to increase the 18k



Rf amplifier section. Q1, Q2 gates are in the center. Note shielding and cutouts for devices.

resistors between pins 10 and 12 and pins 1 and 2 to 22k, which will bring the rolloff to approximately 1 kHz.

step 3. U1 installation. Install U1 and all associated components up to the 0.01/C coupling capacitor from the 455-kHz i-f transformer secondary. During initial test a 0.01/C capacitor can be installed between pins 4 and 9 rather than a filter, and a jumper wire can be added between pin 8 and the output side of the 1k agc feedback resistor from pin 1. Use the PC board terminals provided for this installation. Initially apply power to U1 only and verify that the first section of U1 is working correctly by injecting a 1-mV modulated 455-kHz signal into the input of U1 at pin 2 through the 0.01/C coupling capacitor. By monitoring pin 4, a gain of 25 to 30 dB should be apparent.

table 2. Resonant-circuit component data.

Z1 and Z2	Z3
Inductor: 45 to 50 turns no. 32 AWG (0.2mm) enamelled wire on Miller 4500-2 or 4500-3 coil form	45 turns-same coil form as Z1 and Z2
VVC: Motorola MV1646	None
capacitor: 15 to 20 pF npo or mica type	None
link: 8 to 10 turns, evenly spaced over winding (see text)	same as Z1 and Z2

Fig. 4 is a schematic for the suggested bfo circuit. Initial bfo injection levels should be set at 50 to 75 mV rms; higher values may cause the second section of U1 to oscillate. With injection of both bfo and carrier signals a detected audio signal of 50 to 75 mV should appear at pin 7 of U1. The bfo voltage injection level should be increased to the point of U1 oscillation then decreased slightly. Remove the coupling capacitor between pins 4 and 9 of U1, and install the bandpass filter, using jumper wires from the PC board. Apply power to U1, U2, and U3 and inject the bfo and 455-kHz carrier as before. With 1 mV applied at the input of U1 a clean 90 to 100 mV audio signal should be apparent at U3 output. To simplify initial alignment and test, the bfo frequency should be approximately ±1 kHz from the center of the filter. For instance, using a Collins ceramic filter, I use a

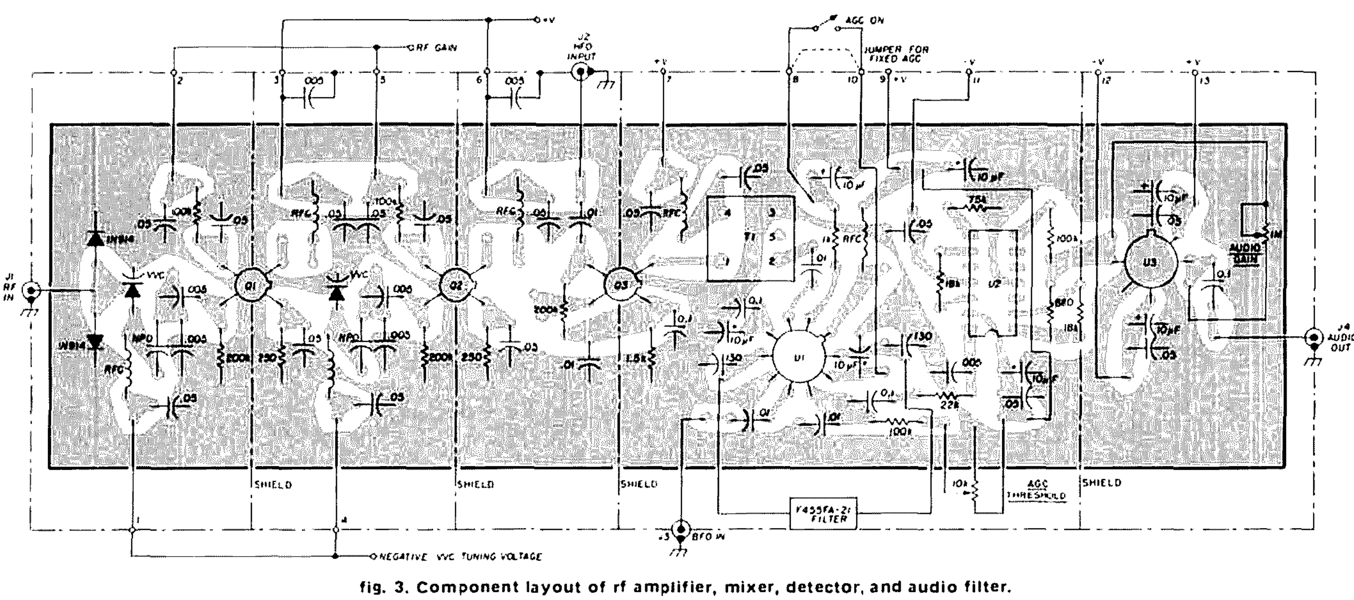


fig. 3. Component layout of rf amplifier, mixer, detector, and audio filter.

455.95-kHz crystal for the bfo frequency, which develops an easily distinguishable 1-kHz audio signal at the output with unmodulated carrier injection.

step 4. Q3 installation. Install all components associated with Q3. Adjust Z3 coil slug for resonance at 3.750 MHz. The input link should be loose along the outside of the coil but terminated to the PC board. Apply power to Q3 and U1 only, and inject a carrier voltage of 3.750 MHz at the link of Z3 and a beat oscillator signal from an external oscillator into G2 of Q3. Reference 1 discusses the design of a suitable high-frequency oscillator that matches this unit. By spreading and sliding the link portion of the Z3 coil input for maximum signal output, a 50-μV signal at Q3 link input should provide a 30-mV detected audio signal at U1, pin 7.

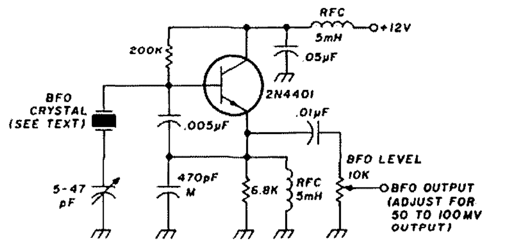
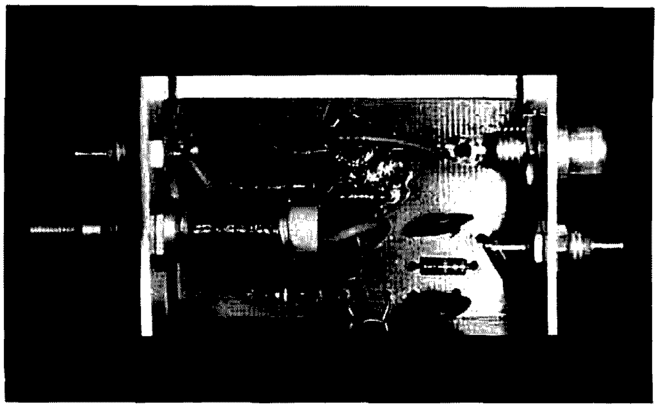


fig. 4. Bfo schematic. Crystal frequency depends on filter, as discussed in text.

step 5. Q2 installation Install Q2 and all associated components except Z2. Apply power to Q2 and Q3 only, and inject a 3.750-MHz carrier into G1 of Q2 while monitoring the drain of Q3. Temporarily apply +12 volts to the input side of the 100-k resistors of G2. Make the final link adjustment on the output of Q2 for a voltage gain of Q2 equal to 8 or 10. Apply a small amount of coil dope to the link to secure it in this position. Install Z2 and the associated VVC components. Apply power to Q2 and Q3 while injecting a low-amplitude 3.750-MHz carrier to the link of Z2. Apply −20 Vdc through 100k pot, and adjust the voltage on the VVC to approximately −10 volts while positioning Z2 slug for resonance. The link may be moved over the surface to the coil to optimize this coupling.



Mixer input section. Q3 gates are on the right; HFO input is through phono jack at upper left.

step 6. Q1 installation. Repeat all installation procedures used for Q2 on Q1. Adjust Z1 slug so that the same corresponding VVC tuning voltage is applied to both resonant circuits in the Q1 and Q2 stages. The operating characteristics of Q1 and Q2 can be verified by alternating grounding G2 of Q1 and Q2 for a short time while monitoring the output of Q3 at either the drain or at the i-f transformer secondary. Temporary grounding of G2 for either Q1 or Q2 should result in an immediate decrease in the output signal level. This should demonstrate the AGC operating feature of the dual-gate mosfet. With the ground removed, the output signal should slowly rise to full value. If you're not interested in using an external rf gain control, you could substitute a single-channel fet for Q1 and Q2 and adjust the source resistance for adequate feedback. Maximum gain is achieved when the source terminals of Q1 and Q2 are at ground potential; however, optimum noise margins are obtained using a source resistance of approximately 220 ohms.

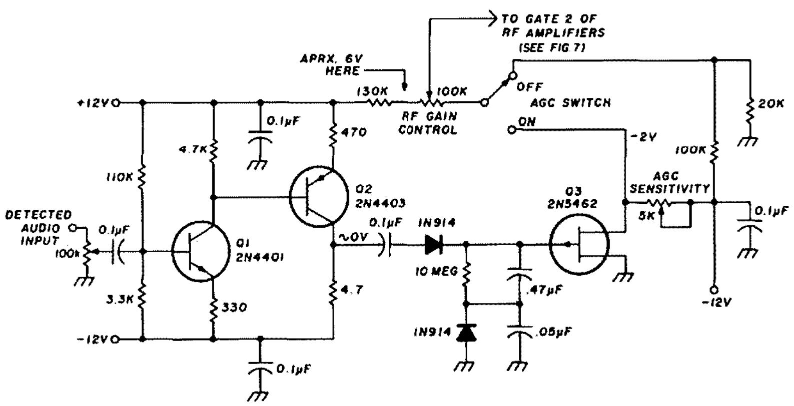
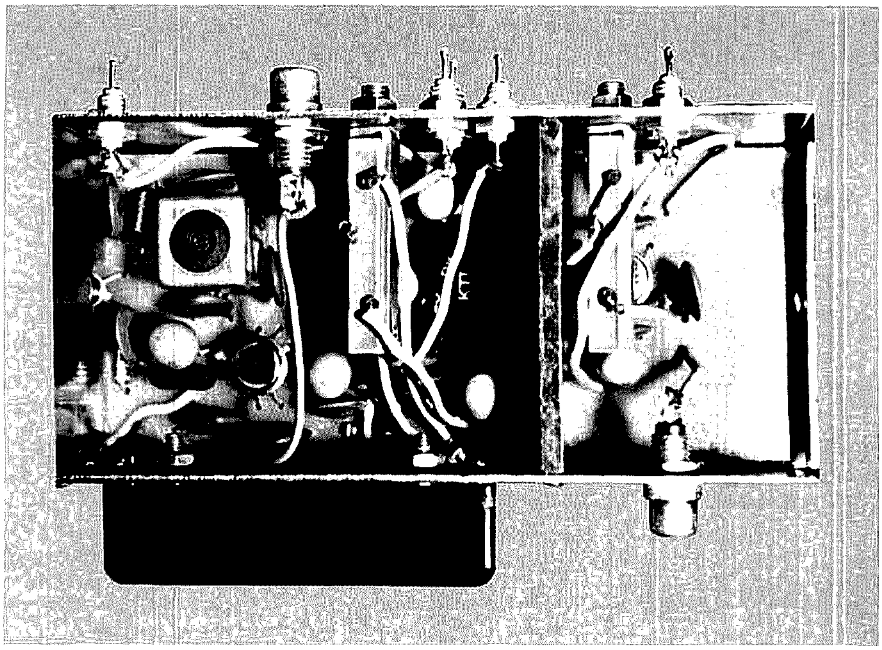


fig. 5. Suggested circuit for agc and rf gain control.



Mixer output, i-f amplifier, and audio filter. Bfo input is through phono jack at right of the 455-kHz i-f transformer. Audio output is at phono jack, lower right.

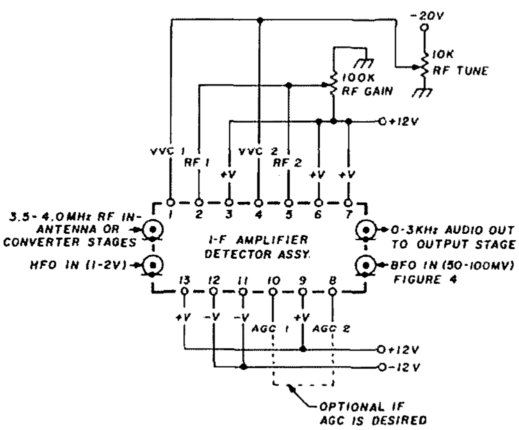


fig. 6. Simplified operating mode for the i-f/detector module.

step 7. Initial i-f testing. During this step if oscillation is apparent, look for feedback caused by the temporary clip leads. Care in placement of clip leads and wire routing to the bfo and hfo is imperative. Improper grounding and bypassing will couple unexpected rf components into the system. As an example, if your bfo is at approximately 456 kHz, then the 8th harmonic is at 3.468 MHz, which can be picked up by the module if you have poor shielding in the high-gain stages.

While keeping the current-limiting resistors in place, carefully apply power to each succeeding stage, starting at U3 and proceeding through Q1. If you haven't made a polarity error, poor positioning of clip leads, or injected too much bfo voltage into U1, the system should be quite stable. Monitoring the output of U3 should display low-level random audio output noise, which should not change even while Q1 input is alternately grounded or left open when the hfo is tuned over the entire input range.

Inject a 5 μV rms signal at 3.750 MHz into Q1 input. While monitoring U3 output with the hfo tuned to the beat frequency, an audio signal of approximately 0.1 V peak-to-peak at 1 kHz should be obtained. Some adjustment in the VVC tuning pot may be necessary to peak the signal.

Strictly speaking, the gain of a system is defined as:

$$dB = 10 \log \frac{|V_2^2|/R_2}{|V_1^2|/R_1}$$

$$\text{If } R_1 = R_2 \text{ then } dB = 20 \log \left|\frac{V_2}{V_1}\right|$$

This relationship is widely misused, and in voltage ratios the gains are expressed in decibels even though $R_1 \neq R_2$.

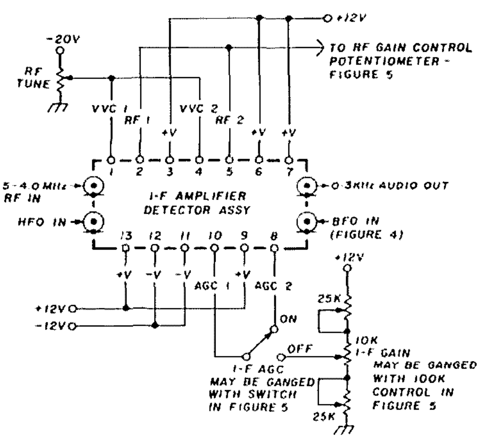


fig. 7. Sophisticated operating mode for the i-f/detector module.

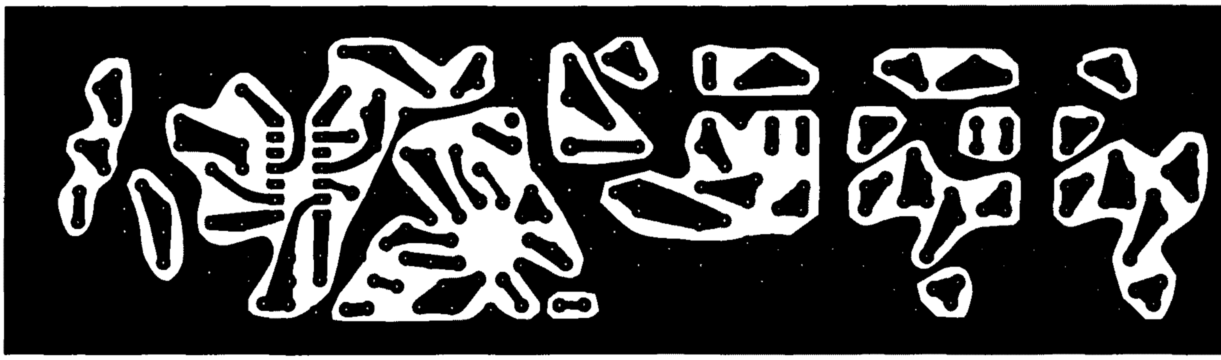


fig. 8. Printed-circuit board layout for the rf amplifier and mixer (left side of board) and the detector and audio filter. Board is 7-7/8" (20cm) long by 2-1/8" (5.4cm) wide.

Furthermore, we had a carrier of 3.750 MHz on the input and are examining an audio signal on the output. If we ignore this small impropriety, we can calculate the i-f/detector gain as:

$$dB = 20 \log \left| \frac{output\ voltage}{input\ voltage} \right|$$

The measured output voltage is 0.1V peak-to-peak, or

$$V2\ rms = \frac{0.1}{2.828} = 0.0353\ rms$$

and the input

$$V_1\ rms = 5(10^{-6})\ V$$

$$dB = 20 \log \frac{0.0353}{5(10^{-6})} = 76.9\ dB$$

final assembly

Remove the temporarily installed current-limiting resistors and install short pieces of bus wire for feedthrough access from the chassis. Install the PC board assembly into the chassis as shown in the photographs, terminating the bus jumpers to feedthrough terminals. Twist the inductors slightly then slide them into mounting holes along the side of the chassis.

For final testing the assembly should be wired as shown in **fig. 6.** Readjust the coils slightly, and if the agc threshold trimmer is used, adjust it for approximately 1 volt. The agc jumper wire between pins 8 and 10 may be removed for initial testing. Apply a 1-μV signal to the input, adjusting the coil slugs and the rf gain and rf tune resistors. A detected 0.1 volt audio signal should appear at the output, which represents a maximum gain of 90.9 dB. The slug settings on Q1, Q2, and Q3 may be adjusted to optimize the tracking characteristics. It won't be necessary to use the full voltage range control from the 10k rf tuning pot since only a few volts are necessary for maximum gain in the rf stages. Peak gain for the mosfet is achieved between 4 to 10 V. Where fixed gain of the i-f/detector is desired, pins 2 and 5 may be tied directly to the +12 volt lines through 100k current-limiting resistors.

By experimenting with the 100k agc-threshold setting, you'll find that a 4 to 5 volt setting on U1 pin 7 will provide a 50% decrease in gain at 300 μV input. You might wish to vary this setting to suit your specific needs. For strictly CW applications this option is not necessary and the circuit shown in **fig. 6** will be easiest to implement. For maximum operation flexibility, the assembly should be operated as shown in **fig. 7.** This arrangement provides agc control at both the rf and i-f stages and, additionally, allows a variable control in the rf and i-f gain to achieve the most ideal signal-to-noise margin in receiving both CW and ssb signals. Referring to **fig. 7,** with the trimpots shown on the i-f gain, a setting range for both the rf and i-f gain control can be achieved that will allow the pots to be ganged, which is also true of the agc switches shown in **figs. 5 and 7.**

With both crystal and ceramic i-f bandpass filters, the ideal ssb bfo frequency is at the plus and minus 20-dB points from center frequency. CW reception is most easily distinguishable at 800 to 1000 Hz. The crystal oscillating frequency then is usually determined after obtaining the filter and measuring the center and 20-dB points. For both upper and lower ssb and CW reception, individual bfo circuits can be used following the design suggested in **fig. 4,** having a common output bus and separate 12-volt control lines so that the desired frequency may be selected.

The selection of Q1, Q2, and Q3 is a matter of preference. The Motorola MFE3006, MFE3008, and RCA 40673 work well in the circuit shown. Recently, RCA announced an economy device, the 40841. Comparison tests between the 40841 and MFE3006 indicate that the RCA device is a comparable unit and could become the receiver builder's most versatile transistor. It may be used as a single-gate device by tying the gates together for most fet oscillator and amplifier applications, or as a general-purpose unit operating to 50 MHz.

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ham radio

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Frequency coverage

1.8 MHz Band	1.8 - 2.0 MHz
3.5 MHz Band	3.5 - 4.0 MHz
7.0 MHz Band	7.0 - 7.5 MHz
14 MHz Band	14.0 - 14.5 MHz
21 MHz Band	21.0 - 21.5 MHz
28 MHz Band A	28.0 - 28.5 MHz
28 MHz Band B	28.5 - 29.0 MHz
28 MHz Band C	29.0 - 29.5 MHz
28 MHz Band D	29.5 - 30.0 MHz
WWV	RX only 10.0 MHz

MIC, input impedance
50KΩ

Frequency
300 - 2700Hz (−6dB)

Sensitivity
SSB less than .25 μv for 10 dB S/N+N ratio
CW less than .15 μV for 10 dB S/N+N ratio

Modes of operation
SSB (LSB or USB)
CW

Input power
200W
1.8-21.5MHz
100W
28-30MHz

ANT, impedance
50Ω - 75Ω Unbalanced

Carrier suppression
More than 50dB

Side band suppression
More than 50dB

Spurious and harmonic suppression
Greater than 40dB

3rd order distortion products suppression
Greater than 30dB

IF Frequencies
1st IF 9MHz
2nd IF 50kHz

Selectivity
SSB 2.4kHz (−6dB)
4.0kHz (−66dB)
CW 400Hz (−6dB)
1.8kHz (−66dB)

Audio output into 8Ω load
2.5W (10% distortion)
3.0W (MAX)

Power source
AC 120V 50/60 Hz (can be re-wired for 240V)

Power drain
400VA TX
78VA RX
48VA RX (Power tube OFF)

Semi-conductors

Transistor	98
IC	43
Diode	120
Tube	3
Digital Ind.	1

Weight
44 lbs. 6 ozs. (23kg)

Dimensions
16-3/4" x 7" x 13-5/8" (420 x 172 x 340mm)

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an introduction to slow-to-fast-scan television converters

Digital scan converters are a new innovation in slow-scan television — here's how they work

One aspect of amateur radio that appears to be leading all others in technical advancement is slow-scan television. Unquestionably, the accelerated advancements in this field are due to the application of digital techniques to circuit design. Most digitized sstv units are also available on PC boards. This means that an enthusiatic amateur can build highly sophisticated sstv circuits without a full knowledge of how they operate. With these thoughts in mind, this article is presented as a guide to understanding one of sstv's latest innovations: the slow-to-fast-scan converter. I'd like to stress that this is an explanatory rather than a construction article. Amateurs can build kits; I'm attempting to describe how their circuits operate – in this case, the slow-to-fast-scan converter.

If you haven't kept abreast of sstv evolution, here's a thought worth considering: begin with a technical starting point (such as this article), expand from there, then follow subsequent advancements. This beats following behind new innovations because you're not familiar with in-between areas.

converter memories

Dynamic mos shift-register memories are the heart of all scan converters. The typical converter memory has about 16,000 digital words by four-bit plane capacity. This means that about 64,000 bits can be stored in such a memory. The four-bit planes usually store from binary 0000 to 1111, or 16 shades of gray. A simplified example of this type memory is shown in **fig. 1**. The top left corner of this hypothetical unit is shown storing a 5. Its binary equivalent is written into the corresponding four-bit planes. Similarly, the binary equivalent of 8 is written into its corresponding planes, and zeros are written into the four right-hand planes. The number of digital words showing on the memory's front area indicate video resolution capability, while the number of bit planes determine gray levels.

If you can visualize 1800 memory cubes as in **fig. 1** wired in series, you'll have a good idea of the memory size used in a slow-to-fast-scan converter. Also, gray coding would be used rather than straight binary-coded decimal coding. This gray code is easier to use because only one bit changes state between successive digital words. Now let's consider the overall concept of scan conversion.

slow-to-fast-scan converter

This converter represents a revolution in sstv technology because it allows you to view slow-scan television pictures on a conventional (fast scan) home television set. Although there are presently only two basic designs of this unit,[1,2] their operational concepts follow a definite pattern. A typical slow-to-fast-scan converter is shown in **fig. 2.** Incoming slow-scan television signals are fed to an sstv demodulator and sync separator. These circuits, shown in the dotted lines of **fig. 2,** represent a typical P7 sstv monitor without a cathode-ray tube and high-voltage supply. In fact, a conventional sstv monitor could be used for this part of the converter if desired.

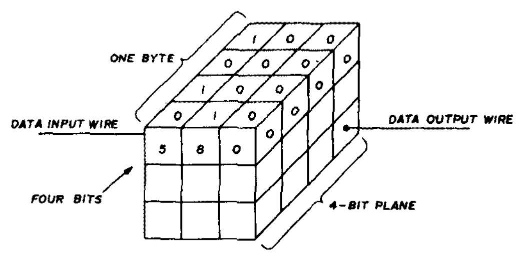


fig. 1. Three-dimensional representation of a shift-register memory used in a slow-to-fast-scan television converter. In this example the total memory capacity is 9 bytes or 36 bits.

Varying video voltages from the demodulator enter the analog-to-digital converter, which changes these voltages to their binary equivalents. This A/D converter employs a voltage divider string connected to voltage comparators to produce digitized equivalents proportional to the varying video levels. This conversion is accomplished periodically by sampling each line of slow-scan video, as illustrated in fig. **3.** The sampling rate is usually 256 times per line. This sampling rate results in 256 words, each containing four bits of gray-code information. Immediately following each line of sstv video, the sync separator extracts a slow-scan sync pulse, which

By Dave Ingram, K4TWJ, Eastwood Village, 604N, Route 11, Box 499, Birmingham, Alabama 35210

triggers the slow-speed clock (fig. 2). This clock, in turn, opens the proper line buffer input to allow each digitized line to enter the 65,000-bit shift register memory. (Meanwhile, the fast-speed clock directs "read from memory" and "recirculate fast scan information" functions. It does this by sending high-speed shift pulses to the memory and timed, binary-1-level signals to the D/A converter).

this operation for, say, 20 times, you'll see that the red cars will again be on the track. This rather poor analogy is similar to scan-converter memory functions.

concluding remarks

This information has been presented with the hope of enlightening interested amateurs on the basic concepts of digital slow-to-fast-scan conversion. Remember that

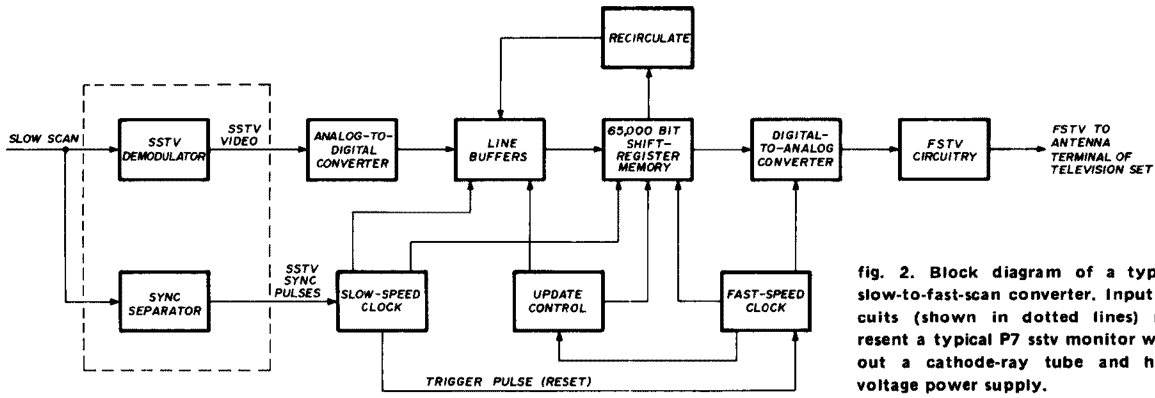


fig. 2. Block diagram of a typical slow-to-fast-scan converter. Input circuits (shown in dotted lines) represent a typical P7 sstv monitor without a cathode-ray tube and high-voltage power supply.

Now, at the precise start of the proper fast-scan line, the fast-speed clock sends a pulse to the update control. This control then directs a binary-1-level signal to the appropriate buffer output, which allows a full line of slow-scan information to enter the fast-speed 65,000-bit memory. This controlled loading operation is performed approximately 128 times during an eight-second period to load a complete sstv picture into memory.

The shift register memory continuously operates at a fast-scan rate, so this information is output (at the proper times) into the D/A converter, where it's converted to fast-scan video. Every 1/15750 second the fast-speed clock sends a sync level pulse to the D/A converter to produce sync-level signals. The D/A converter uses a resistor-transistor network arranged in a *ladder adder* configuration to produce voltage level equivalents to the binary weight of incoming counts.

The final step involves modulating a simple uhf signal generator with the composite fast-scan television signal. This signal is applied to the antenna terminals of a conventional television receiver. Now the resultant slow-scan television picture can be viewed in a well illuminated room.

memory analogy

Memory operation could be illustrated by using an automobile race track to symbolize data flow. First, consider the race track as a circle filled to capacity with very fast blue cars. Fifty green cars are on the entry ramp at one end of this circle, and fifty red cars are on the exit ramp at the opposite end of the circle. A horn sounds and ten red cars quickly leave the exit ramp. At the same time, ten blue cars quickly enter the exit ramp, and ten green cars slowly enter the race track (but quickly gain speed). The red cars then drive around the track and wait in line with the green cars. If we follow

phenomenal progress is being made in slow-scan television technology. While by no means describing the final word in scan-converter design, I think you'll find this article helpful in following the operation of these circuits.

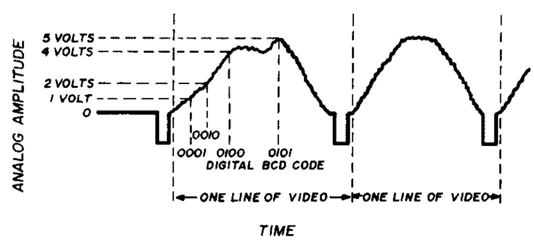


fig. 3. Each line of slow-scan video voltage is sampled at a periodic rate (approximately 256 times per line), which results in 256 digital words per line, each containing four bits of gray-code information.

Other sstv innovations are being developed, such as the color system that uses a 500-Hz subcarrier to carry I and Q color information, while black and white information is contained in the usual 1200 to 2300 Hz range. You'll probably be hearing more on slow-scan television developments in the near future.

I'd like to thank Dr. Robert Suding, WØLMD, for his criticism on the final version of this article. If reader interest warrants, future articles on digital electronics or slow-scan television may be presented.

references

1. David Ingram, K4TWJ, *Slow Scan TV for Amateurs,* TAB Books, 1976, (WØLMD scan converter on page 110).
2. Dr. George Steber, WB8LVI, "SSTV to Fast-Scan Converter," *QST,* Part I, March, 1975, page 33; Part II, May, 1975, page 28.

ham radio

a revealing analysis of the coaxial dipole antenna

This analysis of the coaxial dipole antenna shatters many of the myths — it also explains why its swr bandwidth doesn't measure up to some of the claims

At some time or another most radio amateurs have used a folded-dipole antenna. And, wisely or not, more and more amateurs are now trying the *coaxial* dipole, sometimes called the "double bazooka."[1] Those amateurs who favor the folded dipole generally do so because of its flexible input-impedance characteristics; many amateurs are also acquainted with its somewhat broader bandwidth characteristics as compared with the single-conductor dipole. Many amateurs also try coaxial dipoles in hopes of obtaining increased bandwidth.

Some of the proponents of the coaxial dipole apparently use it because they have been misled into believing it also possesses certain other exotic characteristics, such as supplying its own balun when fed with a coaxial transmission line, as if providing an inherent unbalanced-to-balanced input (not so); exhibiting more than 1.5 dB gain over a simple dipole[2] (not so); and even offering a noise-cancelling capability because the antenna is said to be "shielded" (some amateurs have claimed that "the coaxial dipole has no man-made noise pickup whatever").

One purpose of this article is to show why the coaxial dipole cannot perform according to these utopian specifications and that, except for restricted differences in bandwidth and impedance, its performance is the *same* as that of a simple dipole. Although the coaxial dipole can provide a limited broadening of the impedance bandwidth as compared to a simple dipole *if* engineered correctly, the coaxial dipole configuration in general use by amateurs is not engineered correctly. Consequently, it does not provide the broadband operation erroneously attributed to it.

A second reason for this article is to discuss the *reasons* why the engineering requirements for broadband operation are *not* fulfilled in the amateur configuration, thereby alerting the eager but unsuspecting builder before he wastes valuable time and expensive coax in building a complex dipole that will perform no better than a simple dipole.

impedance bandwidth

The impedance bandwidth of an antenna is derived from the impedance mismatch between the antenna and its feedline as the operating frequency is varied between specified limits. The principal contributor to increased mismatch, as the operating frequency departs from the self-resonant frequency of the antenna, is the reactance which appears in the input impedance of the antenna. The effect of reactance and other parameters is shown in **appendix 1**. This reactance is developed whenever current reflected from the ends of the radiating elements arrives back at the input terminals with other than a 0° or 180° phase relative to the incoming current from the feedline. A zero or 180 degree relation is obtained only when the operating frequency is the same as the self-resonant frequency. This is why the resonant frequency of an antenna is sensitive to the length of the radiating element.[3]

In an *ideal* antenna having infinite bandwidth all the power delivered by the feedline would be radiated by the time the outward-flowing current reaches the end of the radiator, so no reflected current would return to generate a reactance at the input terminals. In other words, the ideal antenna would simply be a broadband transformer, matching the feedline impedance to the 377-ohm intrinsic impedance of free space at *all* frequencies. In our quest for increasing dipole bandwidth we are looking for some scheme which will either cancel or compensate this reactance as it appears (as in folded and

By **Walt Maxwell, W2DU,** Box 215, Georges Road, Dayton, New Jersey 08810

coaxial dipoles), or reduce the reactance by reducing the amount of out-of-phase current reflected from the ends of the radiator. The out-of-phase currents arriving from the ends of a conically-shaped radiator (or a fan-shaped multi-wire dipole) are smaller than those in a thin dipole so less off-resonance reactance is generated in wider-ended radiators.

The reactance compensation obtained in folded and coaxial dipoles results from applying reactance of the opposite kind, contributed by shorted stub sections, as shown in **fig. 1B** for the folded dipole, and in **figs. 1A, 1C** and **1E** for the coaxial configuration. Each half of the folded dipole and the internal portion of each coaxial section in the coaxial dipole forms a resonant quarter-wavelength stub near the resonant frequency of the dipole, short-circuited at the end opposite the feed point. The two stubs of the coaxial dipole are connected in series through their inner conductors. The series combination is shunted across the dipole input terminals as shown in **fig. 1E**. The shunt-connected stub combination reduces the off-resonance reactance appearing at the dipole input terminals because the stub reactance is inductive below resonance, while the dipole impedance is capacitive, and vice versa. Thus, the off-resonance mismatch to the feedline is reduced.

However, the analysis which follows reveals some facts which will probably come as a distinct surprise to many amateurs, and may cause those who are using the coaxial dipole to contemplate replacing it. The facts are, one, whatever increase in bandwidth is obtained solely by the two coaxial stub sections inside the dipole, the *same* bandwidth can be obtained by using a simple wire dipole of the same outside diameter, but having an *external* shunt stub (equivalent to the internal stubs) connected *directly* across the dipole input terminals as shown in **fig. 1D**. Stuffing the stubs inside the radiator does nothing except to provide a convenient place to hide them. This is because the stub currents flowing on the inner side of the coaxial outer conductor are completely separate from the antenna currents flowing on the outside, and the outside antenna currents are *unaffected* whether the conductor is the outer portion of a coaxial cable, or simply a solid conductor.

Secondly, the amount of bandwidth improvement actually obtainable using the shunt-stub technique depends directly on the relationship between the values of the conductance term of the dipole admittance and the characteristic impedances of both the feedline and the shunt-stub lines. This relationship, which will be explained, involves conversion between equivalent series and parallel circuits. It limits feedlines to those having impedance values, Z_c, within a range whose lower limit is well *above* those commonly used (50 or 75 ohms). The stub lines must have Z_c values in the range from five to ten ohms, which are practically unattainable.

These requirements warrant an explanation which will follow shortly. But first, for the reader who is using a coaxial dipole fed with a fifty-ohm transmission line, how about trying an experiment which will prove that the stubs don't provide the heralded broadbanding? Measure and record the vswr at regular frequency intervals across the entire band. Then open the center conductor of the coaxial sections between the two dipole halves at point A in **fig. 1E**, and replace the antenna to the original height. Now remeasure the swr at the same frequency points and prepare for a shock. I predict that you will find an insignificant difference between the two sets of swr readings.

control of mismatch by R and X

To utilize frequencies in any part of a band, antennas are operated off resonance (except at one frequency).

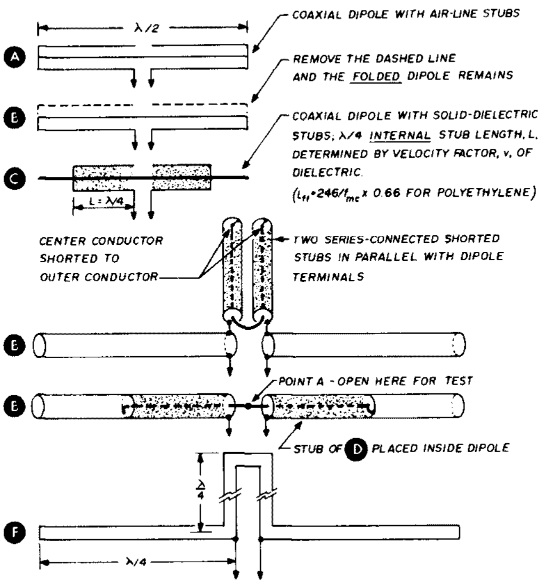


fig. 1. Reactance compensation in coaxial and folded dipoles results from reactance contributed by shorted stub sections. Coaxial dipole with air-line stubs at (A) reduces to a folded dipole (B). Each half of the folded dipole, and the internal portion of each coaxial section in the coaxial dipole (C), form a resonant quarter-wavelength stub. The same bandwidth can be obtained with a simple dipole which has the same outside diameter by using an external shunt stub across the feedpoint (D). Placing the stubs inside the dipole at (E) does not change bandwidth performance. The equivalent configuration is shown at (F).

When operated off resonance we know that the dipole antenna impedance, Z_d, contains both resistance, R, and reactance, X. To obtain zero mismatch at resonance, we know that the line impedance, Z_c, must equal the antenna load impedance, $Z_d = (R + jX)$, which is $R + j0$. To obtain minimum mismatch (which can't be zero) off resonance, the line impedance, Z_c, must equal the absolute value, or magnitude of the load impedance, $|Z_d|$, i.e.

$$Z_c = |Z_d| = \sqrt{R^2 + X^2}$$

A feedline having an impedance which will vary the same as $|Z_d|$ over the desired frequency range does not exist. Therefore a compromise must be found which will now be discussed.

As seen in **table 1** of **appendix 1**, for a value of line impedance which equals load impedance, $|Z_d|$, the

mismatch is smaller when X *is* low *and* R *is high;* mismatch is zero for a load impedance of $R + j0$. We know that the magnitude of the dipole impedance, $|Z_d|$ rises ***above*** its resonant impedance on ***both sides*** of resonance because of the off-resonant reactance component which appears in the dipole impedance (**fig. 2**). I will show later in the analysis that, when applying the stub-compensation technique, the reactance of the stub shunting the dipole *tends* to cancel the off-resonant dipole reactance.

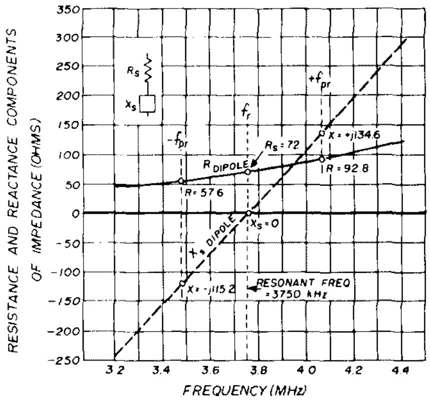


fig. 2. The series resistance, R_s and reactance, X_s components of uncompensated dipole impedance versus frequency. Values shown here are for a half-wavelength thin-wire dipole in free space.[11]

In our stubbing operations the following relationship between equivalent series and parallel circuits is of great importance.

When a series circuit containing both resistance and reactance is shunted with a reactance of the opposite kind, the absolute value of the circuit impedance $|Z|$ is increased. However, in applying the technique of shunt-stubbing to thin half-wavelength dipoles, the transformation between series and parallel circuits appears to have been ignored in amateur literature and applications. The impact of ignoring this important relationship is evident when you see that, while the shunt reactance of the stub *is* cancelling the dipole reactance, the stub reactance simultaneously raises the series-circuit resistance in the *resulting parallel circuit.*

Therefore, even though the effective reactance in the load impedance is lowered by the stub reactance, the resulting increase in resistance increases the magnitude, $|Z|$, of the load to values which are even higher than those of the off-resonant impedance of the dipole alone. Thus you have simply exchanged a mismatch caused by reactance for a similar mismatch caused by the increased resistance.* These relationships are illustrated in the following example:

Consider an 80-meter dipole antenna at a height yielding a resonant impedance, Z_d, of $55 + j0$ ohms at 3750 kHz. At 200 kHz below resonance (3550 kHz) the impedance Z_d is $50 - j90$ ohms. Here $|Z_d| = 103$ ohms, and the mismatch is 5.04:1 on a 50-ohm line (see **appendix 2** for calculation method). Since stubbing is a shunting operation, we now examine the equivalent parallel circuit. The component values of the equivalent parallel circuit are $R_p = 212$ and $X_p = -117.8$ ohms. If the capacitive 117.8 ohms were entirely cancelled, the new load impedance would be $212 + j0$ ohms, and the mismatch would be reduced only to 212/50 = 4.24:1.

Note that the poor swr improvement is because the resistance has been raised from 50 to 212 ohms. However, the situation is even worse if the reactance compensation is performed with RG-58/U stubs generally found in amateur coaxial dipoles. Two 50-ohm RG-58/U stubs in series, resonant at 3750 kHz, yield an inductive shunting reactance of 1190.9 ohms at 3550 kHz. If they are shunted across the above parallel circuit of $R_p = 212$ and $X_p = -117.8$ ohms, the resulting values are $R_p = 212$ and $X_p = -130.7$ ohms. The equivalent series impedance is now $58.4 - j94.7$ ohms, $|Z_d| = 111.3$ ohms, and the mismatch is 4.9:1. Obviously neither of these two mismatch reductions is worthwhile.

However, a stub can provide broadbanding *if it is* properly engineered. The approach is clarified if we recall how mismatch is produced by a constant value of $|Z_d|$ when its components X and R vary if line impedance, Z_c, is nearly identical to dipole impedance, $|Z_d|$: low mismatch when X is low and R is high, as shown in **appendix** 1, table 1. However, in the example above the dipole impedance $Z_d = 55 + j0$, and the line impedance, $Z_c = 50$ ohms, are nearly identical *at resonance.* Thus the mismatch is lowest at resonance: $Z_d/Z_c = 55/50 = 1.1:1$. But as we depart from resonance the minimum obtainable mismatch increases for either of two reasons: The uncompensated dipole impedance, $|Z_d|$ becomes increasingly higher than the line impedance, Z_c, or the parallel-circuit resistance, R_p, becomes increasingly higher than the line impedance. These reasons explain why, in the example, so little mismatch reduction is obtained with stubbing, even with the reactance entirely cancelled. In other words, if there is a substantial difference between the line impedance and the absolute magnitude of the load impedance, it makes little practical difference in the amount of the resulting mismatch whether the load is predominantly resistive or reactive.

On the other hand, as will be shown later, you can reduce the mismatch (using stubs) over a limited frequency range by choosing a line impedance, Z_c, intermediate between the extreme values of the compensated dipole impedance encountered over the frequency range of interest. Since the magnitude of the complex

*A typical article contributing to the misunderstanding on this point, by failing to appreciate the fundamental reactions resulting from *parallel-connected* circuit elements, may be found in *73 Magazine,* June, 1973, page 80 (John Schultz, W2EEY, "The Double-Coaxial Antenna"). The author's statement that a 50-ohm feedline must be used to feed coaxial-dipole antennas further illustrates his lack of appreciation for the actual principles involved.

dipole impedance rises off resonance (and is raised still further by the shunt compensation), the line impedance required to reduce off-resonance mismatch must be higher than that which yields the best match at resonance. Thus we must accept a compromise in the match at resonance in exchange for an improvement in match at frequencies off resonance.

Because our control over the conductance values for high-frequency dipoles of any practical length and length-to-diameter ratio is limited by nature, the relationships just described yield the following important facts concerning the feeding of coaxial-dipole antennas: A useful increase in bandwidth over a simple dipole can be obtained with a coaxially-stubbed *monopole* over a ground plane fed with a 50-ohm feedline because the antenna impedance at resonance is *less* than the feedline impedance. A similar bandwidth increase results when a thin, balanced-feed, half-wavelength coaxial dipole is fed with a feedline having a Z_c of 100 ohms or more. However, *no* significant increase in bandwidth can be obtained using the shunt-stub technique when a balanced, thin, half-wavelength coaxial dipole is fed with a 50-ohm transmission line, because the antenna impedance at resonance is usually near to, or greater than, the 50-ohm line impedance.

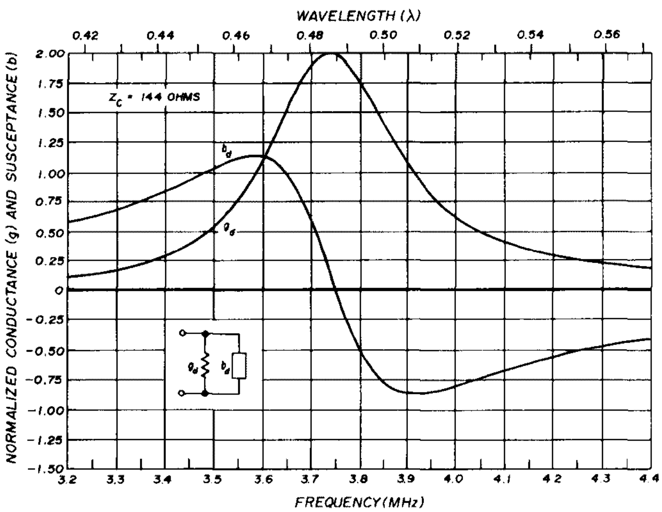


fig. 3. For ease of calculation, the impedance values of fig. 2 have been converted to admittance and plotted in terms of normalized conductance, g, and susceptance, b. Feedline impedance of 144 ohms yields optimum bandwidth, as discussed in the text.

A further requirement for obtaining *significant* bandwidth improvement when using a feedline of suitable impedance is that the shunting-stub reactance drops appropriately with deviation from anti-resonance to compensate the antenna reactance as it increases with deviation from resonance. This requires a stub with a low value of characteristic impedance which is difficult to attain. To satisfy this requirement, the analysis will show that the impedance of the stub-line must be in the range from 10 to 20 ohms if a single external stub is used, or from 5 to 10 ohms each for two stubs *in series*. In typical amateur coaxial-dipole configurations series-connected stubs of RG-8/U or RG-58/U present such high reactance, they are incapable of providing any significant compensation, even if the feedline impedance is of the proper value to obtain *optimum* improvement. Thus two fundamental parameters — low feedline impedance and high stub impedance — cause the typical amateur version of the coaxial dipole to fall short of its goal. The ineffectiveness of these parameters is illustrated in the 80-meter example discussed earlier.

Although this entire situation may seem a bit incredible, the facts are supported explicitly in the published works of Kraus (W8JK),[4] Everitt,[5] Jordan,[6] Coleman,[7] Borton (W9VMQ),[8] and others, in addition to my own experiments and the analysis which follows.* In view of this, how do we account for the apparent success of the coaxial dipole as claimed by so many happy users? There are probably several reasons. First, how many of those happy users performed the test described earlier to obtain a true comparison between the same dipole with and without the coaxial feature? Probably very few, if any. If not, how many amateurs know the *absolute* accuracy of their swr indicators in the *actual* 4:1 or 5:1 swr range? Many indicators give erroneous readings in this range, far *lower* than the true value, resulting in an indication of a wider bandwidth than actually exists.

Secondly, the apparent bandwidth improvement claimed by those using the coaxial dipole with stubs built from 50-ohm cable and fed with a 50-ohm feedline includes the cumulative effects of additional phenomena which are overlooked without realizing that an insignificant amount of improvement is actually contributed by the coaxial configuration itself.

If we assume the swr values presented by Charles Whysall are accurate, they seem to show acceptable results obtained with his Double-Bazooka antenna.[1] However, his design includes *two* broadbanding features which contribute simultaneously — the coax stubs plus capacitive loading of the dipole end sections using two-wire ladder line extending beyond the coaxial portion to obtain a full half-wavelength radiator. This type of multi-wire construction increases dipole capacitance and decreases inductance in the same way as increasing the radiator diameter. It can be shown that this type of loading provides a greater contribution to the increased bandwidth than the coaxial feature, yet Whysall makes no mention of any analysis or experiment performed to determine the *amount of bandwidth* contributed separately by each feature; he simply states that the effective reduction of length-to-diameter ratio provided by the ladderline contributes to a lowered radiator Q.

This is not intended as a criticism of his method of obtaining more bandwidth, or of his article, which con-

*My analysis and experiments were reviewed by *QST's* Associate Technical Editor, Gerald Hall, K1PLP, who performed a verifying experiment at the ARRL Lab. Result: The coaxial-dipole antenna has not been included in the *ARRL Antenna Book.*

tains valuable and pertinent information; it simply isn't correct to attribute the combined effect of *both* features to the coaxial feature alone because it's misleading to the uninitiated.

Since the mechanical construction of the coaxial dipole is not simple, another purpose of this article is to help you to understand what is actually happening so you can reorient your design approach toward a construction method which will yield increased bandwidth performance. Antennas which meet this requirement include the multi-wire, fan-shaped, bow-tie dipole mentioned earlier, invented by Philip S. Carter of RCA in 1937 to obtain the bandwidth necessary for television.[9] Since the beginning of television broadcasting millions of TV receiving antennas have used the Carter bow-tie configuration, and in 1955 it was suggested for 80-meter use by Camillo and Purinton,[10] and again in 1975 by Borton.[8] Additional information on the fan dipole has also been published by Meier.[7]

coaxial dipole principles

In this section we will examine broadbanding principles using stubs under four different conditions. The first two will provide useful bandwidth improvement, while the third and fourth illustrations will provide negligible improvement.

For the analysis to be realistic it will be centered around a half-wavelength dipole which has the physical characteristics of a typical amateur coaxial-dipole antenna, as shown in figs. 1C and 1E. Since the characteristic impedances of both the feedline and the shunt-stub line are critical factors in controlling bandwidth, I will first show how to determine the value of feedline impedance which will yield the maximum possible dipole bandwidth with a given mismatch, or swr, and the impedance required of the shunt-stub line. Next, for comparison, I will determine the bandwidths obtainable (and the stub-line impedances required) when using feedlines of 100 and 50 ohms, respectively. Finally I will show graphically how to assess the bandwidth performance with 50-ohm feedline and 50-ohm stubs.

Since the parallel-circuit relation of the two stubs with respect to the dipole terminals is sometimes difficult to perceive, I will also show why the shorted stub lines are actually in shunt with the input terminals in *both* types of dipoles. Since I will be discussing circuit elements connected in shunt, or parallel, it is convenient to use the less familiar terms of conductance G, and susceptance, B, (components of admittance, Y,) in addition to the somewhat more familiar resistance, R, and reactance, X, components of impedance, Z. Since conductance and susceptance are the reciprocals of *parallel-circuit* resistance and reactance, respectively, they permit the use of direct algebraic addition of the component values associated with each of the parallel-connected elements to determine the total value of the combination. Handling susceptances in this way simplifies the understanding of reactance compensation, especially in the graphical representation in the illustrations.

As an aid to understanding the relationship between equivalent series and parallel circuits, if the series components of impedance are represented by R_s and X_s, and the parallel components by R_p and X_p, then

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \qquad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

while

$$G = \frac{R_s}{R_s^2 + X_s^2} \qquad B = \frac{-X_s}{R_s^2 + X_s^2}$$

The series resistance and reactance components of the *free-space*, uncompensated, thin-wire dipole impedance versus frequency are plotted in fig. 2. This graph shows an impedance of 72 + j0 ohms[11] at resonance.* However, since the effect of elements added in *parallel* with the antenna is best shown on an admittance diagram, the impedance values have been converted to admittance and replotted in fig. 3 in terms of *normalized* conductance, g, and susceptance, b. The normalizing technique will be explained presently. The fundamental relationship between dipole conductance, feedline impedance, and bandwidth will now be described for conditions which yield the maximum possible bandwidth.

As a point of departure, we decide on a maximum practical value of mismatch, or swr, which can be tolerated over the band of interest. We will identify this value by either "swr limit," or "mismatch limit." We will define the band of interest, or bandwidth, as the difference between the frequency extremes where the mismatch limit is reached. As is well known, maximum bandwidth is obtained in the conventional, uncompensated, coax-fed dipole when the feedline impedance, Z_c, closely matches the antenna impedance, Z_d, at resonance. However, as I pointed out earlier, to increase bandwidth we must accept a mismatch at resonance in exchange. In fact, I will show that for any swr limit we may select, we obtain maximum bandwidth by deliberately causing the mismatch to attain the selected limit at resonance. We cause the mismatch to reach the limit at resonance by choosing the line impedance, Z_c, higher than the dipole resonant impedance, $Z_d = R_d + j0$, by the ratio equal to the desired mismatch limit. Thus to obtain maximum bandwidth for any swr limit, the line impedance, Z_c, must be $Z_c = R_d \times swr_{limit}$, where R_d is the value of the dipole resistance at resonance. We will use a mismatch limit of 2 for this entire discussion. Thus the proper line impedance, Z_c, for feeding a 72-ohm

*A free-space dipole was chosen for illustrating the shunt-stub broadbanding principle because its resistance and reactance components of impedance are precisely known. It also avoids complicating the presentation with the effects of mutual coupling with the ground-reflected dipole image at different heights above the ground. Impedances and bandwidths obtained with actual earth-oriented dipoles will therefore differ slightly from the values presented here. The bandwidth of an uncompensated 80-meter dipole at typical heights of 0.25λ, or less, with a 50-ohm feedline will be slightly wider than that of the free-space dipole because the resistance of the earth-oriented dipole at resonance is reduced from 72 ohms to some value closer to 50 ohms, due to the mutual coupling with its image. However, the percentage bandwidth *improvement* obtained when using shunt stubs with earth-oriented dipoles will *not* differ significantly from that presented herein using the free-space dipole data.

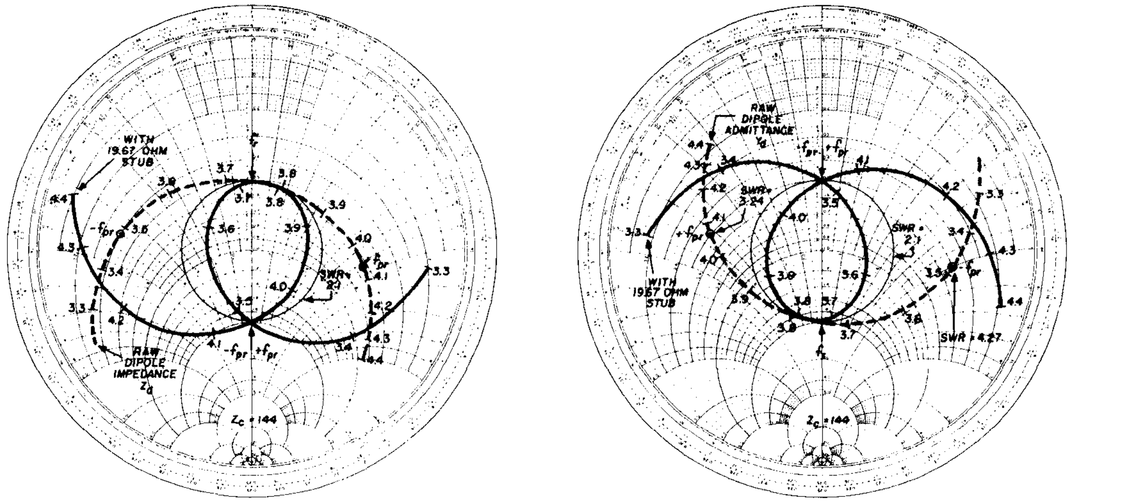
stub-compensated dipole to obtain maximum bandwidth over 2:1 swr limits is $Z_c = 72 \times 2 = 144$ *ohms.*

I will now explain the normalizing procedure and conversion between resistance and conductance, which we will be using in the analysis which follows. An impedance, Z, is normalized by dividing it by the line impedance Z_c, to which it is being referenced; it is indicated by the lower case, z. In our example the normalized resonant-dipole impedance $z_d = r_d + j0 = R_d/Z_c = 72/144 = 0.5$. Conductance, G, is the reciprocal of resistance, R, so $G_d = 1/R_d = 1/72 = 0.0139$ *mhos* (*13.9 millimhos).* Normalized conductance, $g_d = Z_c \times 1/R = Z_c \times G$. Thus the normalized resonant dipole conductance, $g_d = 144 \times 1/72 = 2.0$.

These relations are sometimes best appreciated when displayed on Smith charts, because, in general, plots appearing on a Smith chart represent the normalized values of impedance or admittance. **Figs. 4A** and **4B** show Smith chart plots of the normalized dipole impedance, z_d, and admittance, y_d, respectively; they show conditions both with and without stub compensation. Because of their reciprocal relation these impedance and admittance plots are seen to be mutually inverted. The circle marked "swr = 2:1" encloses all impedances (or admittances) which give rise to mismatch values less than the selected 2:1 limit. These plots will be explained later on.

Since our analysis will be using admittance parameters, the following rule is convenient for determining maximum bandwidth obtained with stub matching in relation to dipole conductance, line impedance, and mismatch limit: The optimum line impedance, Z_c, must be chosen (as explained above) so that the normalized maximum dipole-conductance value, g_d, is equal to the mismatch limit. The maximum conductance, g_d, equals 2 at the dipole-resonant frequency, f_r, as seen in **fig. 4B**; similarly the minimum dipole impedance, z_d, at resonance equals 0.5, as seen in **fig. 4A.**

Since maximum g_d occurs at resonance this establishes the center frequency. The line impedance chosen establishes the band edges of the mismatch limit at the frequencies on either side of resonance where g_d equals the reciprocal of the mismatch limit (where $g_d = 0.5$). Application of this rule in the analysis will be presented shortly. However, it can be seen in **fig. 4** that this relationship between Z_c and g_d is chosen for the compensated dipole to yield an identical mismatch value at the center frequency and at the band edges.



A IMPEDANCE

B ADMITTANCE

fig. 4. Smith chart plots of normalized impedance (A) and normalized admittance (B) from 3.3 to 4.4 MHz for both uncompensated and stub-compensated dipoles, based on the use of a 144-ohm feedline. Selection of 19.7 ohm stubline impedance is discussed in text.

Fig. 4 also shows that selecting the line impedance, Z_c, at 144 ohms gives rise to the 2:1 limiting mismatch at resonance, and places the normalized uncompensated dipole-impedance (or admittance) locus on the graph so it intersects the 2:1 swr-limit circle at resonance (f_r). The importance of this line-impedance selection will be more fully appreciated after an explanation of the reactance-cancelling action. However, it can be seen that placing the uncompensated impedance (or admittance) locus on the graph as just described allows the maximum length (frequency range) of the locus to be warped inside the swr-limit circle as the result of stub compensation. Note that although the mismatch reaches the swr limit at both band center and band edges (because the compensated locus passes through the swr-limit circle at these points), the mismatch is less than the limit everywhere between the center and the edges. This is because all impedances represented by the stub-warped locus lying inside the swr-limit circle give rise to mismatches that are less than 2:1.

The importance of selecting proper line impedance is further emphasized in **fig. 10.** Observe the dramatic reduction in frequency range of dipole impedances warped into the swr-limit circle using 100- and 50-ohm

lines shown in fig. 10, in contrast with the range obtained with the optimum 144-ohm line (fig. 4). The impedances plotted in fig. 10B illustrate the ineffectiveness of stub compensation when using 50-ohm feedline, with either an optimum stub, or 50-ohm RG-58/U stubs. The frequency range of impedances moved into the swr-limit circle with compensation is nearly the same as with no compensation. Thus with 50-ohm feedline, the only conceivable purpose a stub can serve is to enlarge the antenna profile as an announcement to the neighbors, "here is a ham, a potential source of RFI."

This early comparison of the plots in figs. 4 and 10 was simply a preliminary view of the results obtained by examining the four broadbanding conditions using stub-reactance compensation. So we will now return to the explanation of how these results were obtained. It is convenient to use a graph containing dipole admittance data as a worksheet in determining the stub parameters and the bandwidth obtained through compensation using various feedline and stub combinations. We will use the Smith admittance graph of fig. 4B as a worksheet a little later, but we will now use the rectangular-coordinate graph of fig. 5, because it affords graphical construction advantages that assist in clarifying the principles of stub matching. Fig. 5 replots fig. 3, the dipole conductance, g_d, and susceptance b_d (normalized to the optimum 144-ohm line), plus other parameters which will be explained as we proceed.

Using fig. 5, the band edges of the 2:1 mismatch are determined from the dipole conductance curve, g_d, by finding the frequencies on each side of resonance (shown at f_r) where the normalized conductance is 0.5 (the *reciprocal* of the mismatch limit 2.0). We then find a shunt stub of proper *impedance to completely cancel* the dipole susceptance, b_d, simultaneously at both of these frequencies. As seen in both figs. 4B and 5, complete cancellation of dipole susceptance at each band-edge frequency leaves a normalized, pure conductance of 0.5 at both of these frequencies, as required to reduce the mismatch value to the 2:1 limit. The normalized compensated-dipole resistance, r_d, at the band edges is 2.0, as previously stated, and shown in fig. 4A. Thus the real resistance at the band edges equals $r_d \times Z_c = 2 \times 144 = 288$ *ohms*. This reemphasizes that the resistances 72 ohms (at the center frequency) and 288 ohms (at the band edges) yield the selected 2:1 swr limit when they terminate the optimum 144-ohm feedline. Note that this value of line impedance, which yields the maximum bandwidth with a 2:1 swr, is centered between 72 and 288 ohms; it is thus the geometric mean value between the resonant, band-center dipole resistance and the band-edge circuit resistance. With these values and procedures established, we can delve into the operation of the shunt-stub configuration to see how the reactance compensation is effected.

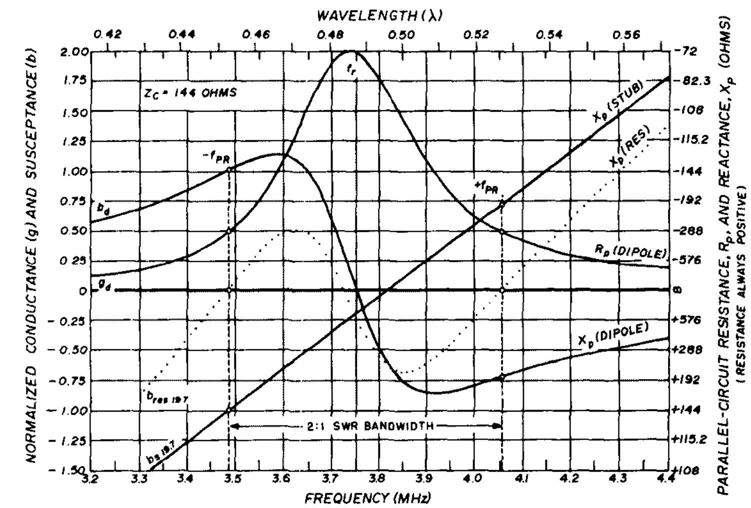


fig. 5. Normalized admittance parameters of a half-wavelength, shunt-compensated dipole. Conductance, g_d and susceptance b_d of the dipole itself are the same as in fig. 3. Since the stub susceptance, b_s, tunes out the dipole susceptance, b_d, at the upper and lower parallel resonant frequencies $-f_{pr}$ and $+f_{pr}$, the resultant susceptance b_{res}, is zero at these frequencies.

shunt stubs

Referring now to fig. 1F, you can see the terminals of *the half-wavelength dipole (shortened to resonance)* connected in parallel with a *shorted* quarter-wave stub transmission line. It is well known that the impedance at the input terminals of a shorted quarter-wavelength line of any characteristic impedance, Z_c, is a pure resistance of a very high value. Thus, at the resonant frequency, f_r, the stub connection has a negligible effect on the 72-ohm dipole impedance. However, at frequencies below resonance both the dipole and shorted stub line become *electrically* shorter, and the dipole becomes capacitive, while the stub line becomes inductive. Due to the parallel connection the inductance of the stub line tends to cancel the capacitance of the dipole.

Conversely, at frequencies above resonance the dipole impedance becomes inductive and the stub line becomes capacitive so a similar compensation is again obtained. Unfortunately, the compensation is far from perfect because, although the dipole and stub susceptances are of opposite polarity, they are not equal nor do they change at the same rate. You can see this in fig. 5 which shows how the dipole susceptance, b_d, varies with fre-

quency in contrast to the stub susceptance shown by the straight line, b_s. The resultant susceptance remaining from the imperfect compensation equals the sum, $b_d + b_s$, shown in the curve b_{res}. This resultant susceptance, b_{res}, combined with the dipole conductance, g_d, will yield the locus of the compensated-dipole admittance in fig. 4B, and the corresponding impedance locus in fig. 4A.

Still referring to fig. 5, if you look far enough above dipole resonance you will find a frequency where the dipole and stub susceptances are equal and opposite, and a perfect susceptance compensation is obtained because this is the higher of the two band-edge frequencies at which the stub line was selected to cancel the dipole susceptance. These equal and opposite susceptance values, $b_d = -0.725$, and $b_s = +0.725$, are seen at points on the ordinate line which intersects the dipole conductance curve at $g = 0.5$, and the frequency scale at $+f_{pr}$ (f_{pr} = *parallel-resonant frequency*). Since the stub susceptance, b_s, tunes out the dipole susceptance, b_d, at this frequency, the resultant susceptance, $b_{res} = 0.0$, and parallel resonance is established. (These values may also be seen in fig. 4B.) From network theory we know that when parallel resonance is obtained by cancelling dipole reactance with a shunt-stub reactance of the opposite sign, then the relatively-low value of series dipole resistance is converted to the higher value of its equivalent parallel-circuit resistance component.

Thus the impedance at the antenna terminals is a pure resistance of 288 ohms for an swr of 2.0 as stated previously. Fig. 2 shows the raw (uncompensated) dipole impedance at this frequency to be $Z_d = 92.8 + j134.6$ *ohms* ($|Z_d| = 163.5$ *ohms*). This impedance would yield a 3.24:1 swr on the 144-ohm line in the absence of the stub. However, the *equivalent parallel-circuit* components are $R_p = 288$ and $X_n = +198.6$ *ohms*, but with the susceptance cancelled, leaving the 288 ohms of pure resistance as shown in both figs. 5 and 6.

Similarly, a second parallel-resonance frequency, $-f_{pr}$, will be found *below* the dipole resonant frequency where the dipole conductance again equals 0.5. This can

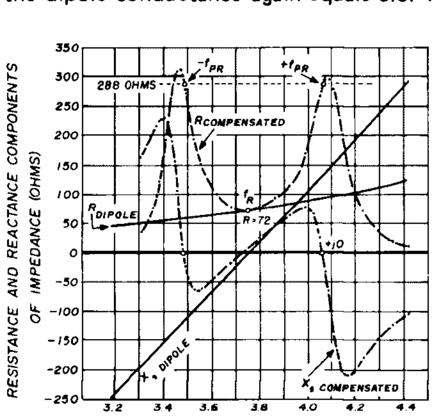


fig. 6. Resistance and reactance components of compensated and uncompensated half-wavelength antennas. Reactance is cancelled at the upper and lower parallel-resonant frequencies $+f_{pr}$ and $-f_{pr}$ leaving a resistive component of 288 ohms.

be seen in figs. 4B and 5. This is the *lower* band-edge frequency at which the stub was selected to cancel the dipole susceptance. The susceptance values at this frequency are $b_s = -1.0$, and $b_d = +1.0$, while the raw dipole impedance shown in fig. 2 is $Z_d = 57.6 - j115.2$ *ohms*, ($|Z_d| = 128.8$ *ohms*) for an swr of 4.27:1 without the stubs. The equivalent parallel-circuit components are

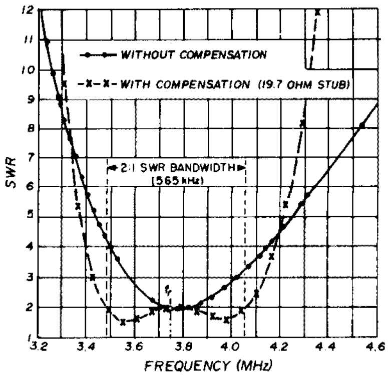


fig. 7. Swr vs frequency for compensated and uncompensated dipoles fed with a 144-ohm feedline. Characteristic impedance of compensating stub is 19.7 ohms.

$R_p = 288$ and $X_p = -144$ ohms, and as at $+f_{pr}$, with susceptances and reactances cancelled, the circuit impedance is again $288 + j0$ ohms for an swr of 2.0:1.

The procedure for calculating the precise value of the shunt-stub line impedance, Z_c, which cancels the dipole susceptance at both band-edge frequencies, from the slope of the stub susceptance line b_s, in fig. 5, requires more space than is available here, but sufficient accuracy will be obtained by using the procedure outlined in appendix 3. However, the Z_c value for this slope is 19.7 ohms and must be divided by two if two stubs are used in series.

We know the mismatch and impedance values at the band center and at the band edges. So we now want to determine impedance and mismatch in the frequency range *between* the band center and the edges. Susceptance, b_{res}, (and reactance) are present in this frequency range, because the stub provided complete susceptance cancellation only at the band-edge frequencies. Determining impedance and mismatch in this range is simplified by using the Smith admittance chart of fig. 4B to calculate graphically the effect of the stub compensation. While we see conductance and susceptance components of admittance plotted separately in fig. 5, these components are plotted together in a single continuous locus on the Smith chart. The component values in the locus are identified separately by the appropriate conductance- or susceptance-circle graduations on the chart grid. We perform graphical addition of stub and dipole susceptances to determine the effect of the stub compensation by adding point-by-point values of the stub susceptance, b_s (obtained from fig. 5) to the corresponding dipole susceptance values (b_d) appearing on the locus of the uncompensated dipole admittance, y_d, on

fig. **4B**. Plotting the results of these additions yields the locus of the stub-compensated dipole admittance. Corresponding values of impedances are found by simply inverting the plots on the admittance graph (fig. **4B**); we thus have plots of the equivalent impedances, as shown in fig **4A**. To show how this works, the normalized, uncompensated dipole admittance is seen to be *0.5 - j0.725* at the upper band-edge frequency, $+f_{pr}$. From fig 5 the stub susceptance, b_s, is *+j0.725* at $+f_{pr}$. Returning to fig. **4B**, we add these dipole and stub susceptances graphically: Following the arrow line along the $g = 0.5$ *circle*, we move point $+f_{pr}$ clockwise for a distance of 0.725 units of normalized conductance, to the chart center where $b = 0$. This move represents the capability of the stub to "pull" or warp the *0.5 - j0.725* point on the uncompensated dipole admittance locus to the new, compensated admittance point, *0.5 + j0*. Going to the corresponding point in fig. **4A**, the admittance-impedance inversion yields the expected normalized impedance of 2.0. The band-edge point used in this example was chosen for simplicity, but the same procedure may be used to convert any point on the uncompensated dipole admittance locus to determine the corresponding compensated value.

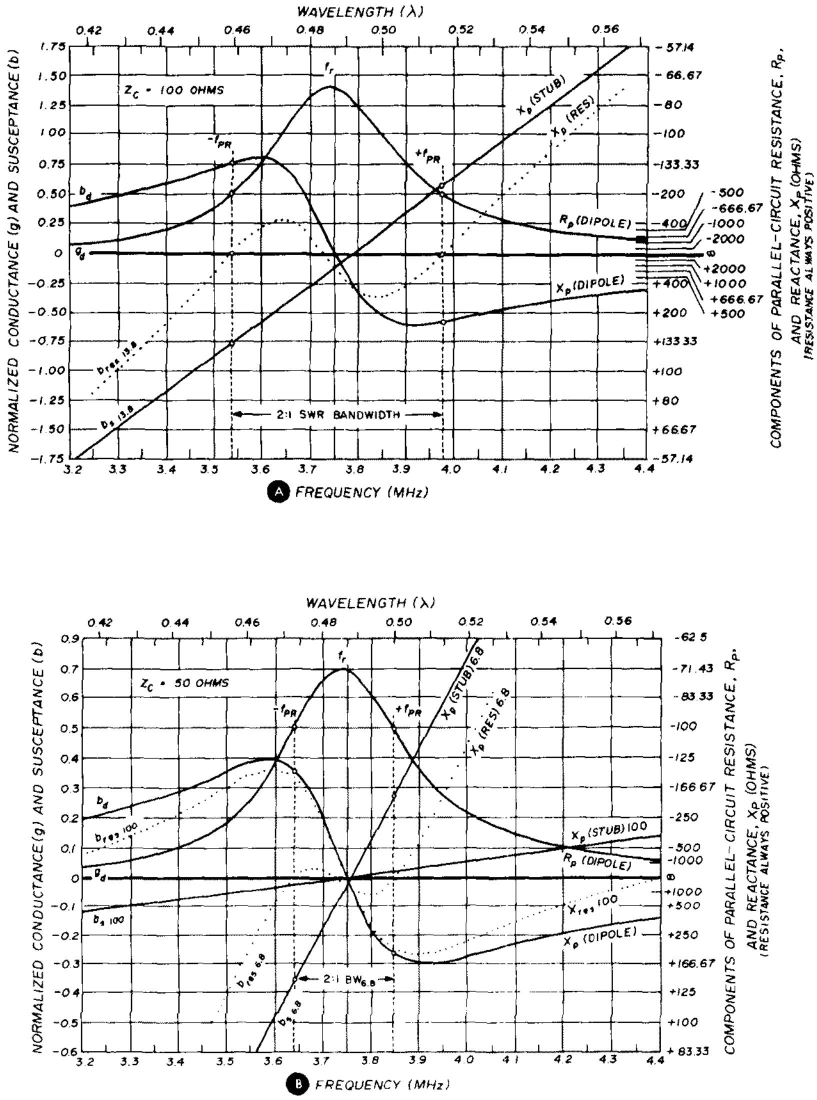


fig. 8. Normalized admittance parameters for uncompensated and compensated half-wavelength dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines. Conductance, g_d, and susceptance, b_d, are for the uncompensated case. Stub susceptance, b_s, and resultant susceptance, b_{res}, are also shown. The optimum stub susceptance for the 50-ohm case (b_s = 6.8 ohms) gives maximum swr bandwidth.

The concept of shunt-stub compensation will now be further clarified by a brief review of some relationships between equivalent series and parallel circuits which form the basis for our operations. These relationships

were stated earlier in the mathematical expressions for converting between series and parallel circuits. This review which follows describes those expressions:

1. Provided Q is greater than 1, the component value of a parallel-circuit resistance, R_p, is higher than its corresponding series-circuit value, R_s.

2. When a series circuit containing resistance, R_s, and reactance, X_s, (an antenna) is shunted by a pure reactance of the opposite kind (a stub), the following changes result:

a. The resistance, R_p, and the conductance, G, components of the equivalent parallel circuit remain constant;

b. The reactance component, X_p, of the equivalent parallel circuit is increased, and the susceptance, B, is decreased.

c. The series-circuit resistance, R_s, is increased, and as the value of the shunting reactance (the stub) is changed in the direction toward cancellation of the net reactance, the series-circuit resistance, R_s, continues to rise until the circuit becomes parallel resonant. At this point the series-circuit resistance, R_s, becomes equal to the parallel-circuit resistance, R_p.*

In view of these relationships, it is important to remember that parallel-circuit resonance exists at both band-edge frequencies because the stub and dipole susceptances cancelled each other to zero. The pure resistance of 288 ohms at the dipole terminals at the band-edge frequencies has been raised to this value because it is the *parallel-circuit resistance* value of the uncompensated dipole impedance, which at both of these frequencies is simply the reciprocal of dipole conductance.

Let's now look at the effect of stub compensation on the separate series resistance, R_s, and reactance X_s, components of the dipole impedance over the entire frequency range extending somewhat beyond the band edges. To observe this effect the dipole conductance and *resultant* susceptance components from fig. 5 are converted into their equivalent series resistance and reactance components of impedance, then plotted in fig. 6 along with dipole resistance and reactance values plotted from fig. 2 for direct comparison with the original, uncompensated dipole impedance components. Note the remarkable change in both the resistance and reactance components which resulted from a change in susceptance *only* – the dipole conductance remained *unchanged* by the stub compensation.

In addition to displaying the bandwidth in the Smith charts of figs. 4A and 4B, the bandwidth obtained with shunt-stub compensation of the half-wavelength dipole in combination with the 144-ohm optimum-impedance feedline is also illustrated in fig. 7 by plotting the feedline mismatch versus frequency. (The mismatch values were computed from the dipole conductance and *resultant* susceptance values in fig. 5 using a technique applicable to pocket calculators described in appendix 2.) For comparison with the bandwidth of the uncom-

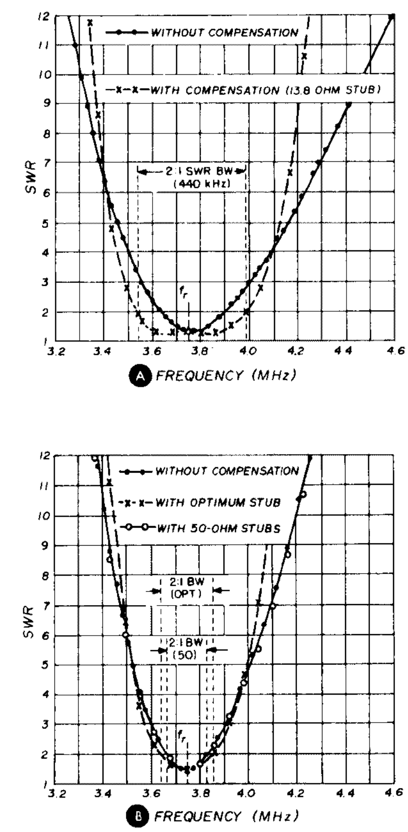


fig. 9. Swr vs frequency for compensated and uncompensated dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines.

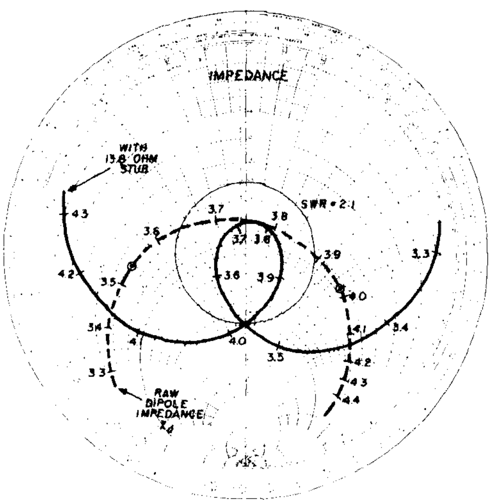
pensated dipole, the mismatch values for the uncompensated dipole were computed from the dipole impedance components in fig 2 and plotted in fig. 7 along with the compensated values.

In contrast to the data shown in fig. 5 for a 144-ohm feedline, figs. 8A and 8B show the effects of using feedline impedances of 100 ohms and 50 ohms, respectively. These are less suitably related to the dipole impedance values encountered over the frequency range than those using the optimum-bandwidth 144-ohm feedline impedance. These graphs show the corresponding normalized conductance and susceptance relationships in the same manner as the 144-ohm case shown in fig. 5. Figs. 9A and 9B illustrate the bandwidths obtained using 100- and 50-ohm feedlines, both with and without compensation. As in the Smith-chart plots of figs. 10A and 10B, they show clearly how the bandwidth drops as the line impedance is reduced from the 144-ohm optimum to 50 ohms.

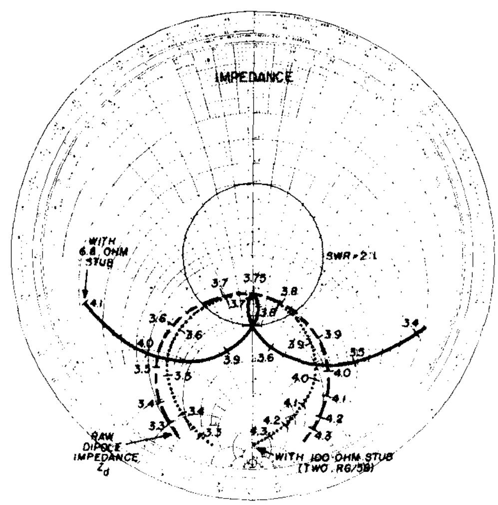
*The basis of the hairpin match used in Yagi arrays. The 20-ohm (approximate) driven element is shortened to introduce enough series capacitance to raise the equivalent parallel-circuit resistance to 50 ohms. The resulting series capacitive reactance (–24.5 ohms) is then cancelled by the hairpin shunt-stub inductance, leaving the feedpoint impedance at 50 + j0 ohms.

From this graphical data you can see why the bandwidth decreases as feedline impedance is reduced. We compare resistance data in **figs. 5, 8A** and **8B** at the 2:1 mismatch bandedge frequencies, where the normalized dipole *conductance is* 0.5. *It is seen that the* parallel-circuit resistance component of the dipole impedance produces the 2:1 mismatch in each case of the three line impedances; the resistance at the 0.5 conductance point being equal to twice the line impedance. This mismatch-conductance relationship holds for any value of feedline susceptance available with a 100-ohm stub, as obtained with two series-connected 50 ohm stubs made from RG-8/U or RG-58/U coaxial cable.

The low slope of this plot is a clue to its meager compensating capability, vividly emphasized by the almost negligible difference between the raw dipole susceptance, b_d, and its corresponding resultant susceptance b_{res100}. Here the uncompensated 2.33:1 mismatch at the $-f_{pr}$ point is reduced only to 2.28:1 with the 50-ohm stubs. The bandwidth increase over the



A Z_c = 100 OHMS



B Z_c = 50 OHMS

fig. 10. Smith chart plots of normalized impedance for compensated and uncompensated half-wavelength dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines. Optimum stub impedance for 100 ohm case is 13.8 ohms; optimum stub impedance for 50-ohm feedlines is 6.8 ohms.

impedance. Therefore, because the parallel-circuit resistance decreases with the use of lower values of feedline impedance, the bandwidth decreases between the two frequencies where the values of equal resistance appear on either side of resonance. Since these dipole-resistance values determine the frequencies of the 2:1 mismatch points, this demonstrates that the maximum obtainable 2:1 bandwidth decreases as the feedline impedance is reduced from 144 ohms.

Let's return to fig **8B** for a closer look at the 50-ohm feedline case. This graph contains susceptance plots of two different shunt stubs with 6.8 and 100 ohms impedance respectively, along with their corresponding *resultant* susceptance plots. The straight line labelled $b_{s6.8}$ plots the susceptance of the stub which completely cancels the dipole susceptance at the 2:1 mismatch points, as shown in the corresponding resultant susceptance plot, $b_{res6.8}$. The uncompensated mismatches of 2.33:1 and the 2.34:1 at $-f_{pr}$ and $+f_{pr}$, respectively, are thus reduced to 2:1. (Not a very significant reduction.) While the impedance of this shunting stub is $Z_c = 6.8$ ohms, it requires a characteristic impedance of 3.4 ohms if two stubs are used in series (not a practical value of Z_c). The bandwidth is increased from 165 kHz (no stub) up to 210 kHz. On the other hand, the plot labelled b_{s100} shows the compensating no-stub, 165-kHz width is negligible; it is even too small to show graphically in **fig. 9B**. The Smith chart plot of **fig. 10B** verifies the disappointing performance of the coaxial dipole when it is fed with 50-ohm line — with *either* 6.8- or 100-ohm stubs.

This analysis of the 50-ohm feedline case, in addition to the mismatch graphs (**fig. 9B**) and the Smith chart impedance plots, clearly shows that there is no significant bandwidth improvement when feeding a coaxial dipole with 50-ohm feedline, especially when 50-ohm coax is used for the shunt-stub lines. Unfortunately, even the optimum 144-ohm line presents nearly insurmountable problems for amateur use. It is true that 144-ohm line could be built using two 72-ohm coax lines in a series, balanced relation — with the outer conductors tied together at both ends, and each inner conductor feeding one dipole half. This would be fine, except that at 80 meters the dipole resistance is usually less than 72 ohms. Unless we find a feedline having an impedance equal to *twice* the resonant dipole resistance, $R_d + j0$, the maximum 2:1 swr bandwidth will not be obtained. Another problem is that determining the optimum feedline impedance ($Z_c = 2R_d$) is simply an academic exercise unless its entire 2:1 swr range of impedances is transformed to the corresponding nominal 50-ohm range required by most amateur transmitters. A

possible solution to this problem may be a broadband transformer described by Jerry Sevick, W2FMI,[12] but probably not without some loss in bandwidth. However, the most severe problem of all is the required low stub impedance — less than 20 ohms. This would require a balanced configuration of two series connected stubs, each less than 10 ohms. Their construction would involve an unwieldy combination of series-parallel quarter-wavelength sections of 50- and 75-ohm coax. Further, the positioning of such a kluge with respect to both its supporting members and the feedline presents additional complications which require more space to explain than is available here.

Since the mismatch values associated with the 50-ohm feedline case in fig. 9B and 10B are higher than those measured by many coaxial dipole users, it is tempting to assume that this analysis is incorrect. However, the following factors responsible for lower *measured* values should be kept in mind.

1. The mismatch values plotted here are those which appear at the antenna-feedline junction. Mismatch values measured at the input end of the transmission line will be lower than at the antenna because of line attenuation (the greater the attenuation, the lower the input mismatch).*

2. In many cases swr indicators read lower than the true value.

3. The use of any additional broadbanding feature such as multiwire end loading, or larger radiator diameter when using RG-8/U coax as the radiator, reduces the inherent reactance which is developed, thereby lowering the mismatch.

Let's now consider WA9PIV's assertions concerning the gain and self-balun characteristics of the coaxial-dipole antenna. Antenna gain is obtained by adding the far-field radiations from each, separate element of any array consisting of more than one dipole element. The coaxial-dipole antenna is not an array, but a single dipole element, and thus has the *same* radiation pattern and the *same* gain as a simple dipole.

Regarding the self-balun characteristics, the bazooka formed by the shorted quarter-wavelength coaxial skirt surrounding a coaxial feedline, as shown in fig. 11, does indeed achieve a balanced-to-unbalanced (balun) action, resulting in cancellation of radiation from the feedline; this would otherwise occur as a result of current on the inside surface of the outer conductor flowing around the top and down the outside of the outer conductor in an admittance path which is in parallel with that half of the dipole fed by the outer feedline conductor. The term "bazooka" cannot be applied to the coaxial-stub configuration within the dipole. As a result, the balun function of the bazooka has been wrongly and unwittingly attributed to the coaxial feature of the coaxial-dipole antenna. The coaxial-dipole antenna is strictly a balanced-input device, and as stated in the opening paragraph, it is the same as the simple wire dipole, except for its impedance and bandwidth characteristics.

WA9PIV's further assertion that all harmonics are rejected by the coaxial dipole is not true because only the even harmonics are rejected. The reason is that the shorted stubs are multiples of a half-wavelength on even multiples of the fundamental frequency, and odd multiples of a quarter-wavelength on odd multiples of the fundamental. Thus a short circuit is reflected across the

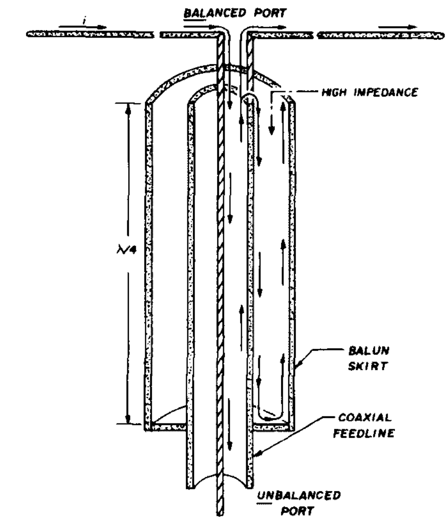


fig. 11. True bazooka formed by shorted quarter-wavelength coaxial skirt surrounding a coaxial feedline provides balanced to unbalanced transformation. This is not the case for the misnamed "bazooka antenna" which is strictly a balanced-input antenna. The high impedance (low admittance) appearing across the open end of the λ/4 resonant cavity of the bazooka shown here impedes current flow into the cavity, thus admitting practically all of the feedline current into the low impedance (high admittance) of the dipole. Removal of the coaxial skirt increases the admittance of current along the outside of the feedline.

input terminals to energy appearing at frequencies of the even-numbered harmonics, while an open circuit is reflected at odd-harmonic frequencies. As a result, harmonic energy is not suppressed at frequencies which are odd-harmonics of the fundamental frequency. These principles also hold true for the folded dipole.

Incidentally WA9PIV specifies dimensions for his coaxial dipole which make the stubs resonant at a frequency higher than the resonant frequency of the dipole. This simply shifts both band edges to frequencies slightly higher than those occurring when the stubs and dipole both resonate at the same frequency. However, regardless of his statement that mismatch is not greater than 1.5:1 over the entire 3.5 to 4.0 MHz band using his stub dimensions, the overall bandwidth is substantially the same as that shown in figs. 9B and 10B.

It is clear, in listening on the amateur bands, that many amateurs, in their attempts to reduce swr, are afflicted with the coaxial-dipole syndrome. It is also clear that many are still placing unwarranted emphasis

*Instructions for calculating the corresponding mismatches appearing at opposite ends of the transmission line, for any value of line attenuation, are presented in appendix 4.

on low swr, usually for the wrong reasons. However, response to my series of articles on this subject[13] indicates that many amateurs are learning that when low-loss feedlines are used (which is generally the case on the high-frequency amateur bands), a low swr *on* the feedline does not save any significant amount of power, nor does it noticeably enhance the level of the radiated signal.

The only significant benefit of a low standing-wave ratio over an entire amateur band is the ease of matching your transmitter output to the *input* to the feedline anywhere in the band. Obtaining an acceptable impedance match is not too difficult on 40 meters, but on 80 meters the line input impedance varies so widely across the band (unless unknown losses reduce the mismatch at the input), it is possible to match the transmitter directly to the line only over a very limited frequency range. Unfortunately, hope for curing this ailment by using the coaxial dipole has been shattered, and anyone dispensing it as a cure should study this analysis.

Following are two suggested prescriptions for an effective-cure of the mismatch problem. As a partial cure try a dose of either W1SX's[10] or W9VMQ's[8] bow-tie dipole configuration mentioned earlier. This antenna provides some improvement over the thin-wire dipole, though not quite enough to permit coupling the average transmitter directly to the feedline across the entire 80-meter band. For a more complete cure, review the *QST* "Reflections" series (especially parts VI and VII),[13] and discover why the mismatch at the antenna-feedline junction that can't be cancelled by the stubs of the coaxial dipole *can be* compensated by conjugate matching at the input terminals of the feedline. Once this is understood, you can live with a three-, four-, or five-to-one swr on the feedline, and still make your transmitter happy by using a tuner to transform whatever impedance appears at the line input to 50 ohms of pure resistance at the tuner input at *any* frequency in the band.

Recommended out-patient treatment for this cure: One hour of operating per day for three or four days using this technique. This treatment will provide sufficient therapy to warrant discharge of the patient, and to guarantee a complete and permanent cure of the coaxial-dipole antenna syndrome.

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appendix 1

The magnitude of an impedance mismatch may be determined as swr from the relationships R/Z_c or Z_c/R *only* when the load is a pure resistance. When the load is a complex impedance, $Z_L = R + jX$, the exact mismatch may be determined in terms of the complex reflection coefficient (**Eq.** 1 of reference 13, part III) from

$$\bar{\rho} = \rho\angle\theta = \frac{Z_L - Z_c}{Z_L + Z_c} \tag{1}$$

The value of swr may then be obtained from the expression

$$swr = \frac{1 + \rho}{1 - \rho} \tag{2}$$

Swr, however, cannot be determined from any relationship between the feedline impedance, Z_c, and the simple absolute magnitude, $|Z|$, of the complex load impedance. This is evident from **table 1** where it can be seen that different values of load impedance, Z_L, can have the same magnitude, $|Z| = 50$ *ohms*, yet produce different values of mismatch on a given feedline.

Furthermore, it is improper to specify impedance as a pure number such as 50 ohms (as it is often heard on the amateur bands) unless it is in some way implied to be resistive, as when referring to the characteristic impedance of a low-loss transmission line. The proper way to specify impedance is to either use the complete polar form (magnitude and angle) or the equivalent rectangular form ($R + jX$) as shown in **table 1**.

table 1. Different swr values on a 50-ohm transmission line as a result of terminating the line with various values of impedance, Z, (different R and X) which have the same magnitude, $|Z| = \sqrt{R^2 + X^2} = 50$ ohms.

load impedance, Z polar form $\|Z\|$ angle	rectangular form R X	swr value (50 ohm line)
50 $\angle 0°$	50 + j0	1.00
50 $\angle 25.84°$	45 + j21.79	1.59
50 $\angle 36.9°$	40 + j30	2.00
50 $\angle 53.1°$	30 + j40	3.00
50 $\angle 66.42°$	20 + j45.83	4.79
50 $\angle 72.54°$	15 + j47.7	6.51
50 $\angle 78.46°$	10 + j48.99	9.90
50 $\angle 84.26°$	5 + j49.75	19.95
50 $\angle 87.13°$	2.5 + j49.94	39.98
50 $\angle 88.85°$	1 + j49.99	99.99
50 $\angle 90°$	0 + j50	∞

appendix 2

Another method for calculating the exact swr from $R \pm jX$ (or $G \pm jB$), developed by the author after a suggestion by W2KF, and derived from **eq.** 1 in **appendix 1**, is shown in the following steps:

1. Normalize by dividing the complex impedance (or multiplying the complex admittance) by the characteristic impedance of the transmission line, Z_c.

$$\frac{R \pm jX}{Z_c} = r \pm jx \qquad (G \pm jB)Z_c = g \pm jb$$

Note that normalized values are in lower case letters.

2. Find the b term of the quadratic formula

$$\frac{b \pm \sqrt{b^2 - 4ac}}{2a}$$

using values of r and x (or g and b) in the expression

$$b = \left(\frac{x^2 + 1}{r}\right) + r \qquad b = \left(\frac{b^2 + 1}{g}\right) + g$$

3. Calculate the swr from the simplified quadratic formula

$$swr = \frac{b + \sqrt{b^2 - 4}}{2}$$

Note that the a and c terms of the complete quadratic formula in **step 2** reduce to one during the derivation due to the normalizing procedure and can be ignored. The negative root of the discriminant

$$-\sqrt{b^2 - 4ac}$$

is also disregarded.

Example: Determine the swr generated on a 50-ohm transmission line by a load impedance of 40 + j30 ohms.

Normalizing: $r + jx = \frac{40 + j30}{50} = 0.8 + j0.6$

Find the b term: $b = \left(\frac{0.6^2 + 1}{0.8}\right) + 0.8 = 2.50$

Calculate the swr: $swr = \frac{2.50 + \sqrt{2.50^2 - 4}}{2} = 2.000$

This is a very good example because the answer is *exactly* 2.0:1 with no fractional remainder. Other examples that give exact answers (50-ohm lines) are 30 + j40, swr = 3.0:1; and 80 + j90, swr = 4.0:1.

To show that different values of load impedance can yield the same swr, the following complex loads (50-ohm transmission line), will generate an swr of 2.6180:1 – 25 + j25, 50 + j50, 100 + j50, and 130 + j10.

appendix 3

Use the following procedure to determine the approximate value of the characteristic impedance, Z_c, of a shunt-compensating stub line from the slope of the susceptance plot:

1. Let θ equal the length in electrical degrees between $-f_{pr}$ and $+f_{pr}$. Find θ by determining the wavelength difference, $\Delta\lambda$, between $-f_{pr}$ and $+f_{pr}$, then multiply by 360 degrees

$$\theta = (\Delta\lambda)360^\circ$$

2. Let X_{av} equal the average of the parallel-circuit dipole reactances appearing at $-f_{pr}$ and $+f_{pr}$ without regard for the sign.

3. The approximate impedance of the stub line can be found from

$$Z_c \approx \frac{X_{av}}{\tan(90 - \frac{\theta}{4})}$$

Example: Determine the impedance of the stub line from the b_s, susceptance plot in fig. 5 ($-f_{pr} = 3490$ kHz, $\lambda = 0.453$; $+f_{pr} = 4055$ kHz, $\lambda = 0.527$).

$$\Delta\lambda = 0.527 - 0.453 = 0.074$$

$$\theta = 0.074 \cdot 360^\circ = 26.64 \text{ degrees}$$

parallel-circuit reactances: $-f_{pr}$ 144 ohms
$+f_{pr}$ 192 ohms
X_{av} 168 ohms

$$Z_c \approx \frac{168}{\tan(90 - \frac{26.64}{4})} = \frac{168}{\tan 83.34} = \frac{168}{8.56} = 19.62 \text{ ohms}$$

appendix 4

The mismatch SWR_A at the antenna feedline junction is higher than the mismatch SWR_I measured at the input to the transmission line because of line loss. When one mismatch is known use the following procedure to calculate the unknown mismatch at the opposite end of the line. Let

ρ_A = reflection coefficient at antenna (point A)

ρ_I = reflection coefficient at input (point I)

α = line attenuation in dB (multiply dB per foot times length of the line in feet)

r = decimal value of the output/input power-loss ratio of the feedline:

$$r = antilog_{10}\left(\frac{\alpha \text{ in } dB}{10}\right)$$

Example: If the line attenuation is 0.5 dB, what is the output/input power-loss ratio? (0.5 dB is expressed as a negative quantity since it is loss.)

$$r = antilog_{10}(-0.5/10) = 0.891$$

A. Use the following steps to calculate the mismatch at the antenna (SWR_A) from an swr measurement at the input to the transmission line (SWR_I).

1. Calculate ρ_I from SWR_I $\rho_I = \frac{SWR_I - 1}{SWR_I + 1}$
2. Calculate output/input power-loss ratio, r, from line attenuation, α
3. Calculate ρ_A from ρ_I/r (ρ_A is larger than ρ_I)
4. Calculate SWR_A from ρ_A $SWR_A = \frac{1 + \rho_A}{1 - \rho_A}$

Example: The input swr to a 120-foot RG-8/U feedline is 3.5:1 at 4.0 MHz. What is the swr at the antenna? (Attenuation of RG-8/U is 0.32 dB per 100 feet at 4.0 MHz so attenuation of 120 feet is 0.384 dB.)

$$\rho_I = \frac{SWR_I - 1}{SWR_I + 1} = \frac{3.5 - 1}{3.5 + 1} = \frac{2.5}{4.5} = 0.556$$

$$r = antilog_{10}\frac{-0.384}{10} = 0.915$$

$$\rho_A = \frac{\rho_I}{r} = \frac{0.556}{0.915} = 0.607$$

$$SWR_A = \frac{1 + \rho_A}{1 - \rho_A} = \frac{1 + 0.607}{1 - 0.607} = \frac{1.607}{0.393} = 4.093{:}1$$

B. Use the following steps to calculate the swr at the input of the transmission line (SWR_I) from a mismatch measurement at the input to the antenna (SWR_A):

1. Calculate ρ_A from SWR_A $\rho_A = \frac{SWR_A - 1}{SWR_A + 1}$
2. Calculate r from line attenuation, α
3. Calculate ρ_I $\rho_I = \rho_A \times r$ (ρ_I is smaller than ρ_A)
4. Calculate SWR_I from ρ_I $SWR_I = \frac{1 + \rho_I}{1 - \rho_I}$

Example: The swr at the input to an antenna is 5:1 at 4.0 MHz. What is the swr at the input of a 156.25 foot length of RG-8/U transmission line? (Attenuation of RG-8/U is 0.32 dB per 100 feet at 4.0 MHz so attenuation of 156.25 feet is 0.5 dB.)

$$\rho_A = \frac{SWR_A - 1}{SWR_A + 1} = \frac{5.0 - 1}{5.0 + 1} = \frac{4.0}{6.0} = 0.667$$

$$r = antilog_{10}\frac{-0.5}{10} = 0.891$$

$$\rho_I = \rho_A \times r = 0.667 \times 0.891 = 0.594$$

$$SWR_I = \frac{1 + \rho_I}{1 - \rho_I} = \frac{1 + 0.594}{1 - 0.594} = \frac{1.594}{0.406} = 3.926{:}1$$

ham radio

differential keying circuit

Low-cost TTL logic devices are combined in a keyer design with optional weight control

An investigation of differential keying circuits in tube, transistor and relay form revealed that such circuits could be designed using TTL or cmos IC logic. Some IC multivibrators were found in the Texas Instruments *TTL Data Book.*[1] One was the SN 74123, which has two multivibrators in one package, and the other was the SN74121, which has only one. The SN74123 was chosen since this meant only one 16-pin socket, and the second multivibrator could be used for another function such as weight control.

circuit description

Reference 2 indicated that the SN74121, SN74123-series inputs require a clamped threshold near ground or at a slightly negative voltage (about -1.2 volt). I selected a clamp at ground because otherwise another power-supply bias level would be necessary; this clamped threshold at ground was adequate for the keyer application.

Waveforms and a diagram of the circuit are shown in figs. 1 and 2 respectively. By following these two illustrations, we can run through the operation of how the differential keying is formed. The NOR and NAND logic is necessary to develop the gating needed. An SN7402 was selected as the NOR gate (J or N type – the pin-out is different for other types), and the SN7410 was selected as the NAND element simply because I had some left over from a previous purchase from Poly Paks (one of many sources for these devices). **Fig. 2** shows one of the four dual-input NOR gates used as an inverter and the clamp to ground for the input of the SN74123 (U1A-B). Both inputs are connected through a current-limiting resistor to a high logic level, thus producing a low at the output, which is very close to ground (0.8 volt or so). The input biasing by means of R1 also provides bias to my keyer output transistor switching stage.

When the keyer output goes to ground U3C output goes high and U1A, being connected for a positive trigger input, produces the inhibit pulse, P3, seen in **fig. 1A.** The completion of the keyer pulse, P1, results in a negative trigger, which is received by the second half of the SN74123, (U1B). Thus, U1B develops pulse P5. Pulses P2 and P3 are processed by NAND gate U2. The output conforms to NAND logic in that an output will occur only with like inputs (positive logic is used, so this means both the inputs must be high). Therefore pulse P3, being negative, inhibits the output during its time interval, and output pulse P4 is formed.

Now the gate pulse that will keep the oscillator on must be developed. This gate pulse is formed by adding the output of U3C with that of U1B (positive or Q output in this case). The final gate output becomes that shown by P6. The third NOR gate of the SN7402, U3B, is used for another inverter, so a positive pulse, P7, is generated.

interface circuits

We now come to the all-important function of interfacing the developed logic pulses with the equipment being considered. Transistors Q2 and Q3 perform this interface and are general-purpose, high-voltage npn and pnp transistors needed for the hybrid interface between logic IC levels and the usual high bias levels of vacuum tubes. R7 provides limiting for the NOR load current, and Q2 base current. The positive output pulse, P7, switches Q2 on, which in turn switches Q3 on, and the

By **Fred Griffee, W4IYB,** 8809 Stark Road, Annandale, Virginia 22003

transmitter oscillator plate and/or screen receives its positive bias voltage.

Instead of using complete on-off switching of bias voltage or voltages, a resistor should be added between the positive bias and the oscillator circuit being controlled, so that the bias level is decreased but still allows low-level oscillation. When full positive bias is applied, the output increases to its normal level in a smooth transition from low to high levels rather than starting from zero oscillator output, which could result in transients. R8 limits Q3 base current and provides additional isolation should the transistors fail. The output switching transistor collector and emitter can be connected in various ways to satisfy the interface requirement of whatever circuit switching is needed.

The keying-pulse interface is made fairly easy by merely using the same type pnp transistor used in the gate interface and turning it on when the NAND gate output goes on (pulse-switches from a high positive level to a low-level ground). R6 provides the gate load and transistor base-current limiting, while Q1 switches the grid-block circuit to ground, thus keying the transmitter exciter and starting the pulse shaping process. The keying pulse is symmetrically located within the gate time interval and therefore performs the desired differential keying function. Normally, grid-block keying circuits are of a high-impedance level so that no further protection is needed should the transistor fail. If Q1 failed, the exciter would remain keyed to ground, but no harm would befall integrated circuit U2.

weight control addition

If weight control is not included in the keyer (or if a bug or hand key is used), then another multivibrator, such as the SN74121, must be used with an additional NOR gate. In fact, the NOR gate used for forming the acceptance gate must be a three-input type (the SN7427 is available from most sources). The three-input NOR can also be used for the two-input requirement by merely connecting two of the three inputs together (see fig. 3).

The logic of these gate arrangements is most easily understood by describing the different timing waveforms again (see fig. 1B). Input pulse P1 is the keyer or hand-key output and turns off U3C when grounded or when the contacts are closed. This action increases positive

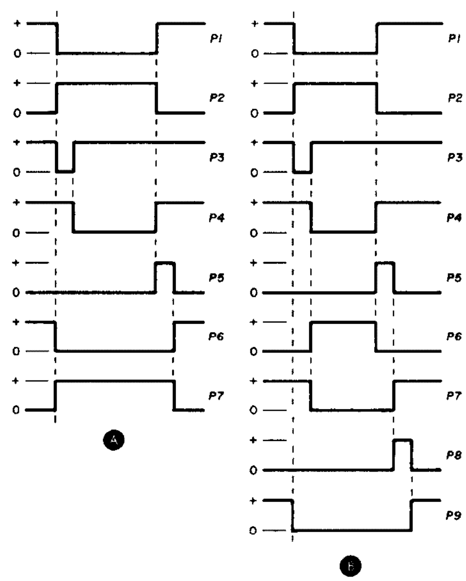
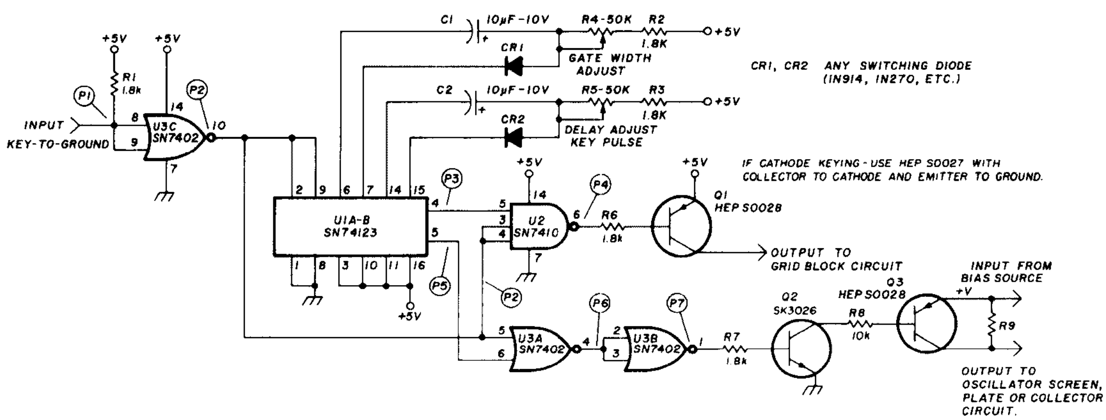


fig. 1. Timing waveforms of differential keyer logic with weight control included in keyer (A) and with weight control added (B).

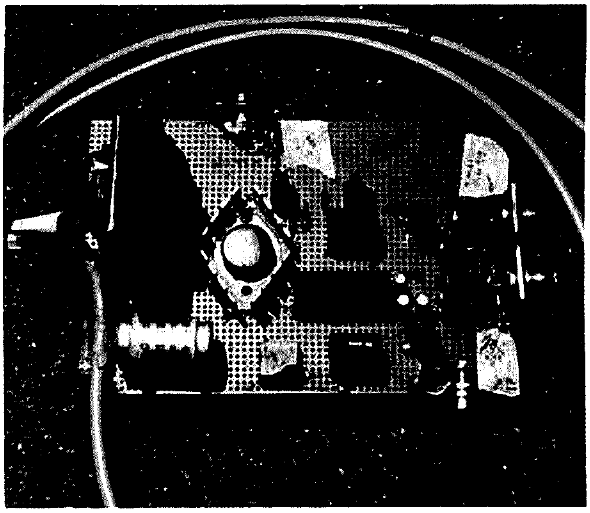
pulse P2 from ground (about 0.8 volt) to about +3.4 volts. One-half of the SN74123 still forms the negative-going inhibit pulse, which forms the keyer output pulse delay, P3, by preventing any output from NAND gate U3A until all inputs are at a high level. The second half of the dual multivibrator U1B (SN74123) is connected to accept a negative trigger input, thus forming the weight-control pulse, P5.

The trailing edge of P5 triggers acceptance-gate pulse



IF CATHODE OR EMITTER KEYING - CONNECT EMITTER TO CATHODE OR EMITTER OF OSCILLATOR AND COLLECTOR TO GROUND. ADJUST R9 FOR NEAR ZERO OSCILLATOR OUTPUT.

fig. 2. Differential keyer schematic without added weight control.



Component side of perf board showing general layout.

stretcher multivibrator U2 (SN74121) and forms P8. Now, P2, P5, and P8 are added by the three-input NOR gate, U4B, producing the required acceptance-gate pulse, P9, that will turn on the VFO or oscillator before and after the keying pulses have been completed (P5 and P6). Keying pulse P4 is inverted by U3B, forming P6, which is then added to the weight-control, multivibrator (U1B) output P5 by U4A, forming the desired keying-pulse interval, P7.

Note that the acceptance-gate interval, P9, is formed by the keying-pulse width, P2, the weight-control multivibrator output, P5, and the gate-extension pulse width, P8, such that any variation in the weight-control pulse also varies the gate acceptance interval, thus inadvertent extension of the keying pulse interval, P7, beyond the acceptance-gate interval, P9 is prevented. Such extension of P7 would cause spurious radiation in the form of key clicks. Also, if any variation is desired, only the weight-control multivibrator time interval adjustment is necessary.

It may be found that when the speed is varied, the weight control may have to be changed slightly. Since exciter or transmitter keying characteristics are usually fixed and drive levels change with frequency, the weight control may have to be changed to obtain the same desired keying quality or characteristic (with class-C amplifiers).

The same low-to-high level interface transistor switching used in the differential key circuit without weight control is used here; however, U3B was required to allow the proper positive level input for U4A with respect to the keyer pulse, P4. U3B merely serves as an inverter by connecting all inputs together.

construction

Nothing is critical in this circuit. Straightforward point-to-point insulated wiring was used. One rf decoupling capacitor was found necessary at the input of the differential keyer circuit (0.001 μF ceramic). My glue gun came in very handy for strapping down sockets and components before wiring. The glue takes a minimum length of time to dry, but be careful about using the glue on temperature-sensitive components since it's initially very hot.

The power supply consists of a filament transformer rated at one ampere (much more than required), an LM309K regulator, and a bridge rectifier IC. All were mounted on the same board. The filter capacitor is a 1500 μF 10V electrolytic. A zener diode would work as well as the LM309K and requires less space.

alignment

Alignment was easily accomplished with the aid of an oscilloscope such as an EICO model 460. Alignment

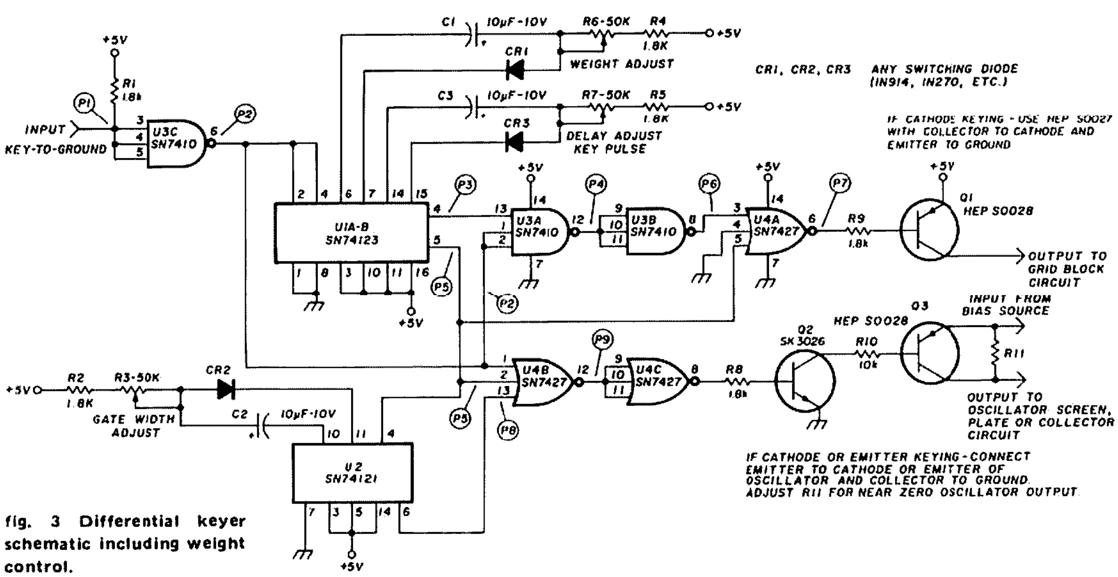
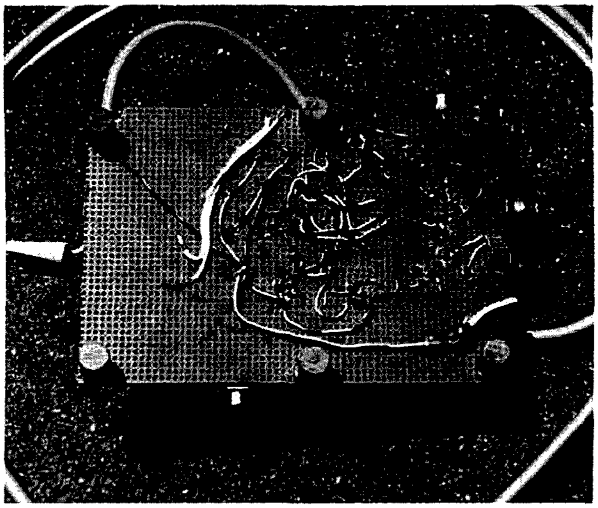


fig. 3 Differential keyer schematic including weight control.

would be almost impossible without an oscilloscope. My Johnson T/R switch has an rf output test point, which allows the transmitted wave shape to be observed when it's within the oscilloscope bandpass, otherwise the scope detector probe can be used and the envelope observed.



Underside of perf board showing point-to-point wiring.

Delay multivibrator U1A pulse width was adjusted by R5, fig. 2, until the front of the shaped pulse started to decay with near zero rise time, then R5 was backed off until the shaped pulse reappeared. This procedure was also used by varying R4 while observing the pulse trailing edge with respect to the acceptance-gate pulse-width stretcher multivibrator, U1B. These controls require no further adjustment, and the keyer weight control is used for any differences that may be experienced with varying code speeds.

If the weight-control logic is added the weight control multivibrator would be set at minimum pulse width, while the acceptance-gate multivibrator pulse width would be varied to eliminate any decay of the shaped-pulse trailing edge. The weight control can then be adjusted to produce the desired keying characteristic.

The circuit has worked well without any malfunctions. Incidentally, the TTL logic has bias limits. It will not work properly or will fail completely if the bias becomes greater than 5.5 volts. A bias value between 4.7 and 5 volts with good regulation by means of a zener or IC regulator device is recommended.

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ham radio

TTL IC tester

IC sockets too expensive? With this tester you can check ICs before soldering them into place

I can remember when you **could** plug a one-dollar vacuum tube into a ten-cent socket. The situation now seems to be reversed: a twenty-cent integrated circuit plugs into a socket costing anywhere from fifty cents to a dollar or more. When I was planning a construction project that would require 63 ICs, sockets were out of the question – just too darned expensive. This simple tester was then constructed so that each IC could be tested before it was soldered into place. The ICs shown were for the project mentioned above. Some could be deleted or other types added according to your favorites or the contents of your junk box.

circuit description

Nine input stimuli are generated by feeding +5 volts through resistors R1 through R9 as shown in **fig. 1**. Each input stimulus is either a 1 (+5V) or a 0 (ground), depending on the position of its corresponding switch S1 through S9. The output states are indicated by lamps DS1 through DS6. If the lamp is ON, a 1 is indicated; OFF indicates a 0 state. Q1 through Q6 can be any npn transistor that will carry the 150-mA lamp current. I used some TO-5 germanium transistors removed from computer PC boards. An alternative output indicator circuit could be an LED with the proper current-limiting resistor in place of the transistor and lamp. The 7413 Schmitt trigger is used as a "de-bouncer," allowing a single pulse to be produced by manually depressing switch S11 (**fig. 2**). This pulse is used in testing JK flip-flops.

construction

The test sockets and lamp drivers are mounted on a piece of perf board approximately 4½ x 2-3/8 inches (114 x 60mm) (**fig. 3**). This board is mounted on 5/16-inch (8mm) spacers above an aluminum chassis, which is 4 x 5 x 1½ inches (101 x 127 x 38mm). All switches and lamps are mounted on the aluminum chassis, and resistors R1 through R9 are mounted underneath. The test sockets are wired in parallel; that is, all pins requiring a no. 1 input are connected together and to S1 and R1; all

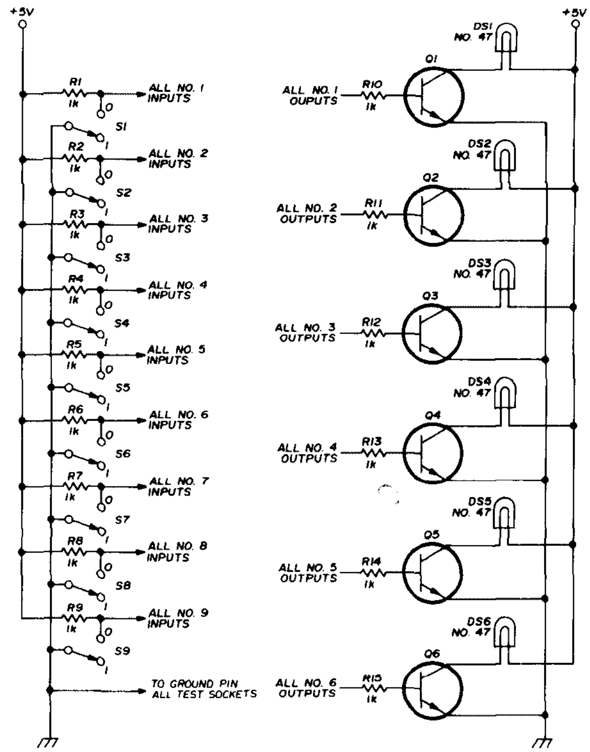


fig. 1. Schematic of the IC tester.

no. 2 inputs pins are connected together and to S2 and R2, etc. A 5-volt power supply at 1 ampere is required to power the IC tester. If you're careful not to light more than two lamps at a time, 0.3 ampere would be sufficient.

operation

The diagrams shown in fig. 2 may be used as a guide when operating the tester. Perhaps, when testing JK flip-flops, it would also be advantageous to have the specifi-

By Kenneth H. Leiner, WA4LCO, 3254 Inverness Court, Orlando, Florida 32806

table 1. Logic matrix for selected ICs.

IC	S1	S2	S3	S4	S5	S6	S7	S8	S9	S10	S11	DS1	DS2	DS3	DS4	DS5	DS6
7400	1	1	1	1	1	1	1	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	1	0	1	0	1	0	1	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
	0	1	0	1	0	1	0	1	NU	NU	NU	ON	ON	ON	ON	NU	NU
	0	0	0	0	0	0	0	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
7402	1	1	1	1	1	1	1	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	1	0	1	0	1	0	1	0	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	0	1	0	1	0	1	0	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	0	0	0	0	0	0	0	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
7404	1	1	1	1	1	1	NU	NU	NU	NU	NU	OFF	OFF	OFF	OFF	OFF	OFF
	0	0	0	0	0	0	NU	NU	NU	NU	NU	ON	ON	ON	ON	ON	ON
7410	1	1	1	1	1	1	1	1	1	NU	NU	OFF	OFF	OFF	NU	NU	NU
	1	1	0	1	1	0	1	1	0	NU	NU	ON	ON	ON	NU	NU	NU
	1	0	1	1	0	1	1	0	1	NU	NU	ON	ON	ON	NU	NU	NU
	0	1	1	0	1	1	0	1	1	NU	NU	ON	ON	ON	NU	NU	NU
	0	0	0	0	0	0	0	0	0	NU	NU	ON	ON	ON	NU	NU	NU
7473	0	1	0	1	1	1	NU	NU	NU	X	YES	OFF	ON	OFF	ON	NU	NU
	1	0	1	0	1	0	NU	NU	NU	X	YES	ON	OFF	ON	OFF	NU	NU
	X	X	X	X	0	0	NU	NU	NU	X	NO	OFF	ON	OFF	ON	NU	NU
	1	1	1	1	1	1	NU	NU	NU	X	YES	TOGGLE		TOGGLE		NU	NU
7476	1	0	1	0	1	1	1	1	NU	X	YES	ON	OFF	ON	OFF	NU	NU
	0	1	0	1	1	1	1	1	NU	X	YES	OFF	ON	OFF	ON	NU	NU
	X	X	X	X	1	1	0	0	NU	X	NO	ON	OFF	ON	OFF	NU	NU
	X	X	X	X	0	0	1	1	NU	X	NO	OFF	ON	OFF	ON	NU	NU
	1	1	1	1	1	1	1	1	NU	X	YES	TOGGLE		TOGGLE		NU	NU

X = Don't-care condition (1 or 0). NU = Not used. TOGGLE = DS1 and DS2 (DS3 and DS4) alternate with each S11 pulse.

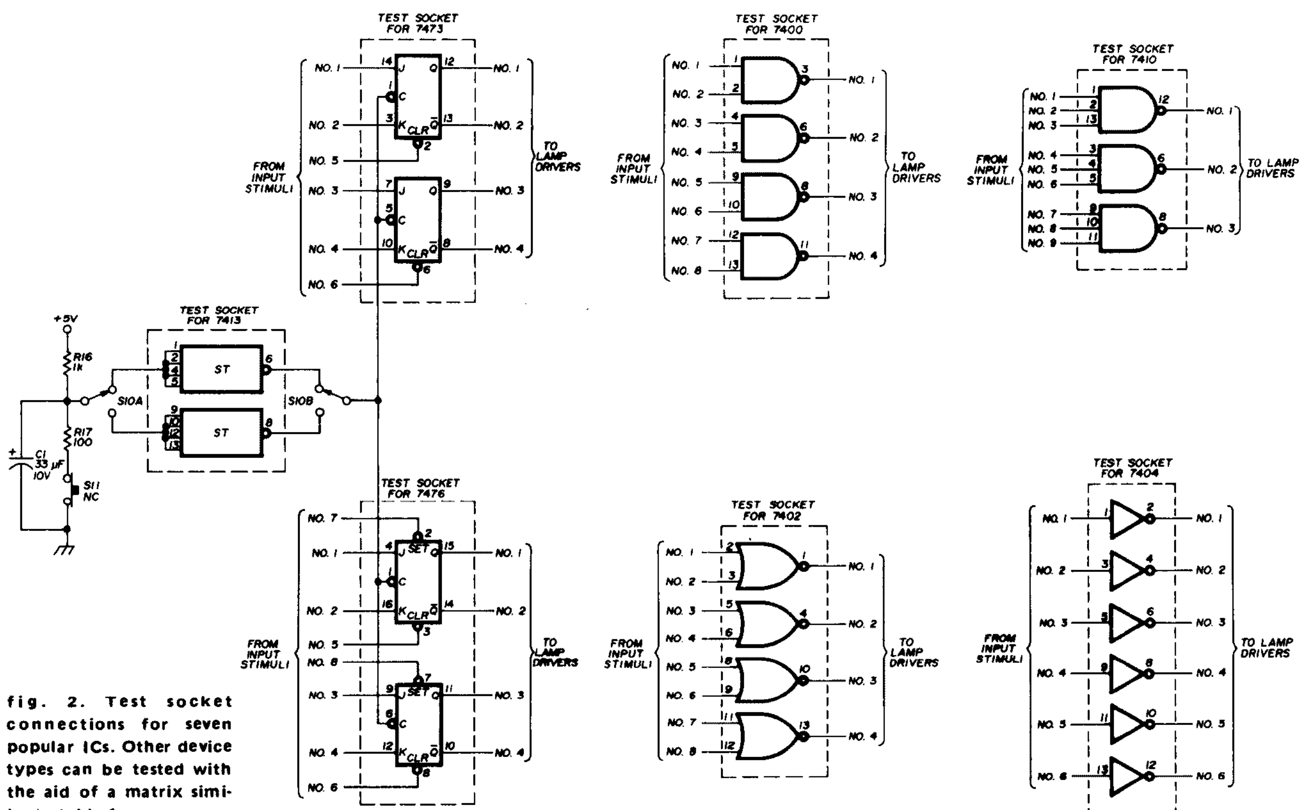


fig. 2. Test socket connections for seven popular ICs. Other device types can be tested with the aid of a matrix similar to table 1.

cation sheet handy. If desired, tables may be constructed similar to **table 1**. Always turn off the 5-volt power when inserting or removing an IC from the socket. In removing ICs from the test sockets, slip a small screwdriver blade under the IC. Then rock the IC gently up and down, raising first one end slightly, then the other end. Repeat until the device is free of the socket. This procedure usually removed the IC without bending any of the metal pins.

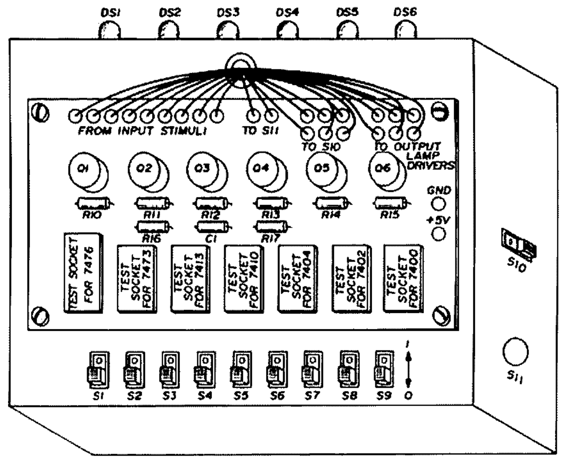


fig. 3. Suggested component layout.

When testing JK flip-flops, a 7413 must also be inserted in its socket. Also, a 7413 can only be tested if a 7473 or a 7476 is in the tester. As wired here, the 7413 has all its input pins connected together, so the test for the 7413 is not complete. The 7413 was really included to produce clock pulses for testing JK flip-flops, so whatever test is available for the 7413 is a bonus. Switch S10 allows both Schmitt triggers of a 7413 to be tested. At all other times, only one IC at a time is inserted into the tester. In testing flip-flops, when power is first applied it is sometimes necessary to depress S11 twice in making the first test. Probably this is because when power is turned on the various internal flip-flops may assume any state, and it takes one pulse to get them into the proper relationship.

conclusion

The tester was used to check about 100 integrated circuits. Three defective units were found. A 7410 had a defective gate; pin 6 remained high at all times. A 7402 showed a dim glow on DS2 and DS4 when they should have been OFF. (This trouble may not have been found if LEDs had been used as output indicators.) Finally, one-half of a dual JK flip-flop remained in the 1 state at all times.

It may be doubtful at times who'll win, but at least when debugging the 63-IC project I had a head start on Murphy and his wretched laws.

ham radio

50-MHz bandpass filter

A bandpass filter of unusual design that provides 6% bandwidth at 50.5 MHz with only 4-dB insertion loss

The design of highpass, lowpass, and bandpass filters for use at hf and vhf has been covered in recent amateur literature.[1,2,3] The professional literature has also offered design aids in the form of slide-rule devices for use in filter synthesis[4] and in graphs.[5] The article on hf bandpass filters for receivers by W7ZOI[3] is an excellent example of showing what can be done and how simple these filters can be. W7ZOI is to be commended for combining amateur know-how with laboratory equipment to demonstrate the selectivity of his designs. His **fig. 7** (reference 3) at first appeared too complex and at the same time reminded me of a similar filter I had hiding in the garage.

filter characteristics

The garage relic is of unknown origin and as **fig. 1** shows, is rather sophisticated. **Fig. 2** is a plot of this filter's response taken from an x-y recorder (using a hand-tuned signal generator). The insertion loss (4 dB) and a bandwidth of 6 percent at the 3-dB points seem pretty good, considering the 50.5-MHz center frequency and the amount of wire on the coils.

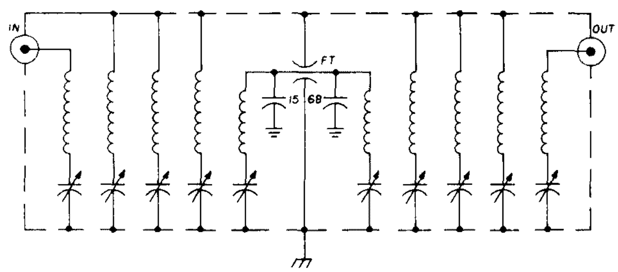


fig. 1. Schematic of the vhf bandpass filter. Center frequency is 50.5 MHz, bandwidth 6 percent, and insertion loss 4 dB. Each of the inductors is about 2.2 μH; variable trimmers are 1.5-7 pF.

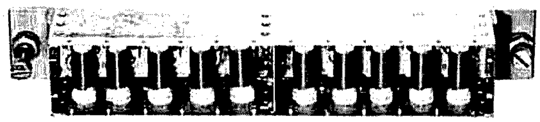
A photo from a Hewlett-Packard spectrum analyzer (**fig. 3**) shows the skirt slope. Vertical divisions are 10 dB and the horizontal scale is 1 MHz/cm.

construction

I don't recommend construction of this filter unless you have a sweep signal generator and a 5-inch (13cm)

By **Paul H. Sellers, W4EKO**, 4002 Columbus Avenue, Norfolk, Virginia 23504

oscilloscope for alignment. Alignment is tedious and quite ticklish. For those brave enough to attempt to duplicate this filter, the photo and fig. 4 are provided. The coils forms are ribbed Teflon rod, ½ inch (12.5mm)



Inside the 50-MHz bandpass filter showing coil arrangement and center shield. (Photo courtesy Paul Ireland).

in diameter. Each of the coils are 21 turns no. 20 AWG (0.8mm) wire; winding length is 1-3/64 inches (26.5mm). Coil ends are inserted through holes in each end of the Teflon rod (fig. 4A). Overall coil diameter,

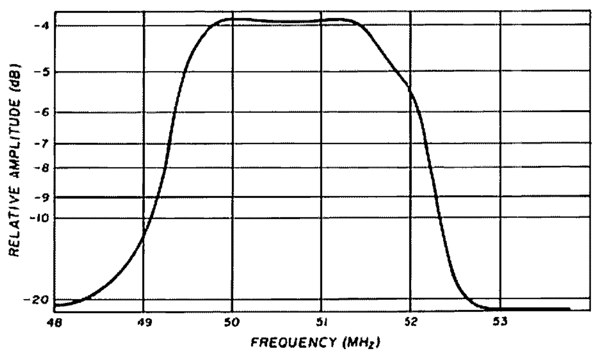


fig. 2. X-Y recorder plot of filter response. Filter insertion loss pushes down the peak of the curve allowing skirts to show out-of-band values of signals passed.

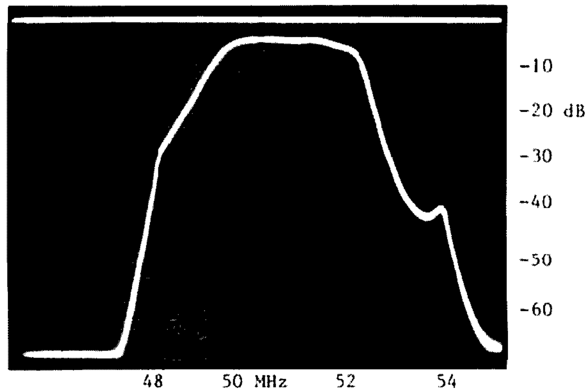


fig. 3. Photo from H-P spectrum analyzer showing filter skirt slope. Vertical scale: 10 dB/division; horizontal scale: 1 MHz/cm.

including wire and ribs, is 9/16 inch (14.5mm). The coils are spaced as in fig. 4B, which is a side view of the filter showing the hold-down screws for the coils. The variable capacitors are about 1.5-7 pF, and the fixed capacitors are Corning type CY10C, 150J and 680J.

I'm sure you'll appreciate the design of this filter, including the unusual input/output circuits and the purely inductive coupling between stages.

references

1. Bob Myers, W1FBY, and Clarke Green, WA1JLD, "Field Day Filter," *QST*, April 1973, page 11.
2. Neil Johnson, W2OLU, "High-Frequency Low-pass Filter," *ham radio*, March, 1975, page 24.
3. Wes Hayward, W7ZOI, "Bandpass Filters for Receiver Preselectors," *ham radio*, February, 1975, page 18.
4. *Genistron Filter Slide Rule*, Genistron Inc., Los Angeles, California, 1965.
5. "Pick a Filter From this Chart," *Electronic Design No. 24*, November 23, 1972.

ham radio

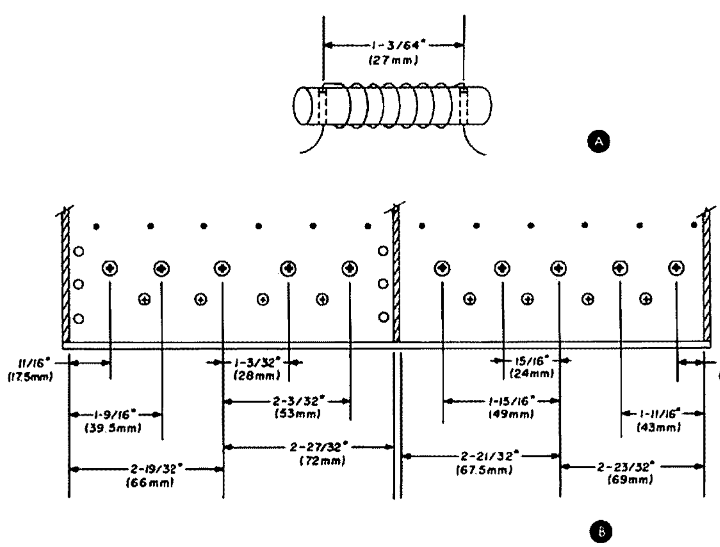


fig. 4. Coil construction details (A) and side view of filter box showing coil spacing (B). Enclosure dimensions are 2-1/8 inches (54mm) deep and 1-7/8 inches (47.6mm) across opening.

Logic Probe 1 is a compact, enormously versatile design, test and trouble-shooting tool for all types of digital applications. By simply connecting the clip leads to the circuit's power supply, setting a switch to the proper logic family and touching the probe tip to the node under test, you get an instant picture of circuit conditions.

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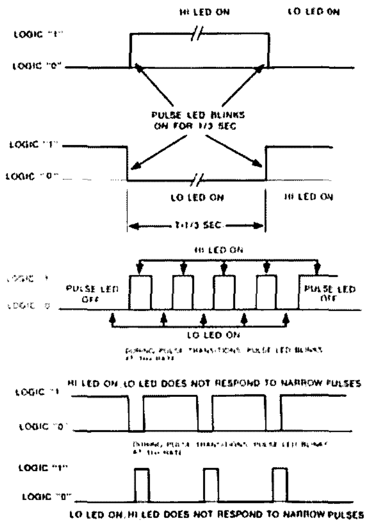

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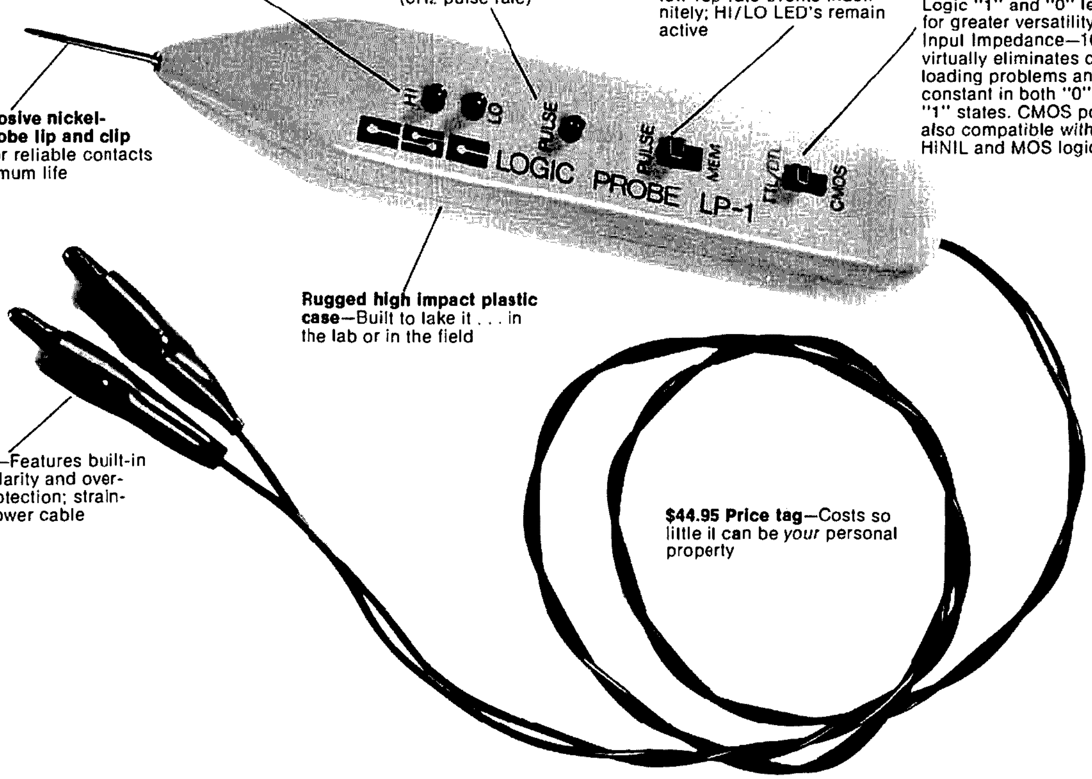

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More Details? CHECK—OFF Page 126

MICROPROCESSORS

microcomputer interfacing: how does a microcomputer make a decision?

One of the most important programming characteristics in any digital computer, including a microcomputer, is the ability to make a decision. For a typical microcomputer, we can define a *decision* as the process of determining further action based on the logic state of a *flag*. A *flag* is a single flip-flop that can be either set or cleared in response to operations occurring within the microcomputer system. A change of state of the flag is usually an indication either that a particular operation has been completed, or that a certain condition exists as a result of a microcomputer operation. Flags can be located either internally or externally to the microprocessor chip; those discussed here are the internal flags, which are set or cleared in response to specific types of microprocessor instructions, such as arithmetic and logical instructions.

The flags located within the microprocessor chip are typically associated with the *arithmetic-logic unit (ALU)*, a region within the chip where all arithmetic and logical operations are performed. In the 8080 microprocessor chip, for example, five flags indicate the following conditions;

By Jonathan Titus, David G. Larsen, WBHYJ, and Peter R. Rony

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

Zero flag If the result of an arithmetic or logical operation is zero, the zero flag is set to logic 1; if nonzero, the zero flag is reset to logic 0.

Sign flag If the result of an arithmetic or logical operation is negative, the sign flag is set to logic 1; if positive, the sign flag is reset to logic 0.

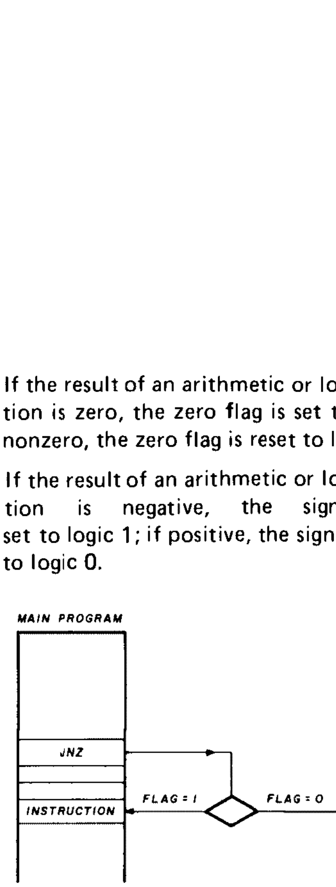


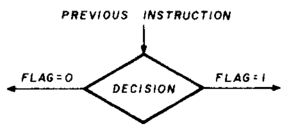
fig. 1. The JNZ instruction. If the zero flag is at logic 1, the instruction is ignored and program control passes to the following instruction.

Parity flag If the result of an arithmetic or logical operation has even parity, the parity flag is set to logic 1; if odd parity the parity flag is reset to logic 0.

Carry flag If the result of an arithmetic or rotate operation has a carry out of the most-significant bit of the 8-bit result, the carry flag is set to logic 1; if not, the carry flag is reset to logic 0. The carry flag is reset to logic 0 after all logical operations.

Auxiliary carry flag If the result of an arithmetic operation has a carry out of bit 3 into bit 4 of the 8-bit result, the auxiliary carry flag is set to logic 1; if not, the auxiliary carry flag is reset to logic 0. The auxiliary carry flag is reset to logic 0 after most logical operations.

Insufficient space is available in this column to discuss all of the above flags, so we shall restrict our attention to the zero flag. Shown below is the traditional flow chart *decision symbol* applied to an 8080 microprocessor decision:



The next instruction executed depends on the logic state of the flag associated with this specific decision. For example, consider the JNZ instruction, where JNZ means "Jump if Not Zero:"

instruction code	mnemonic	description
302	JNZ	If the zero flag is at logic 0, jump to the 16-bit
<B2>		memory address given in bytes <B2> and <B3>
<B3>		of this three-byte instruction; if the zero flag is at logic 1, ignore this instruction and proceed to the following instruction.

The statement "Jump if Not Zero" refers to the 8-bit result of a preceding instruction, not to the logic state of the zero flag. When this result is zero, the zero flag is set at logic 1 and program control passes to the next instruction, as shown in **fig. 1**.

The JNZ instruction is widely used in the creation of programmed *time delay loops,* an example of which is provided in **table 1**. In this program, both the address and instruction bytes are in octal code; it is assumed that the HI memory address byte is 000. The program first moves an 8-bit timing byte into register B; this byte, indicated by an asterisk, has any value between 000 and 377. The value of the byte will determine the duration of the time delay.

At LO memory address 002, a device-select pulse is generated to set the SN7474 flip-flop shown in **fig. 2.** The contents of register B are then decreased by 1.

table 1. Microcomputer program that demonstrates a simple time delay loop based on a decision made on the logic state of the zero flag. This program generates a single output pulse, the duration of which is determined by the timing byte at location 001, at the Q output of the SN7474 flip-flop.

LO memory address	instruction byte	mnemonic	clock cycles	description
000	006	MVI B	7	Move following timing byte into register B
001	*	—	—	Timing byte for register B
002	323	OUT 2	10	Generate device-select pulse that sets the SN-7474 flip-flop
003	002	—	—	Device code for set input to SN7474 flip-flop
004	005	DCR B	5	Decrement contents of register B by 1
005	302	JNZ	10	If zero flag is at logic 0, jump to the memory address given by the following two address bytes; otherwise, ignore this instruction
006	004	—	—	LO memory address byte
007	000	—	—	HI memory address byte
010	323	OUT 3	10	Generate device select pulse that clears the SN7474 flip-flop
011	003	—	—	Device code for clear input to SN-7474 flip-flop
012	166	HLT	7	Halt the micro-computer

*May have any value between 000 and 377. Its value determines time-delay duration.

The JNZ instruction immediately tests the logic state of the zero flag; if the contents of register B are not zero, the flag is at logic 0 and a jump occurs back to LO memory address 004. The DCR B and JNZ instructions are executed repeatedly until the contents of register B become zero, at which time the zero flag becomes logic 1. The JNZ instruction tests the flag for the last time

and shifts program control to the OUT 3 instruction at LO memory address 010. This output instruction generates a device-select pulse that clears the SN7474 flip-flop. Once this has been done, the microcomputer comes to a halt.

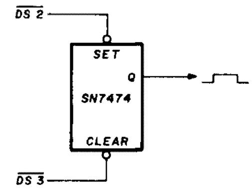


fig. 2. SN7474 flip-flop used as a monostable multivibrator.

The program shown in table1 generates a single output pulse, the duration of which can take any value between 0.0125 and 1.925 ms in steps of 0.0075 ms. Some typical pulse widths are summarized in table 2 for an 8080-based microcomputer that operates at a clock rate of 2 MHz. The calculations associated with the conversion of clock cycles to pulse width were discussed in reference 1. The number of clock cycles is a measure of the actual time it takes the microcomputer to execute a single instruction or group of instructions.

table 2. Examples of output pulse widths generated by the program in table 1 with an 8080 microcomputer operating at a clock rate of 2 MHz.

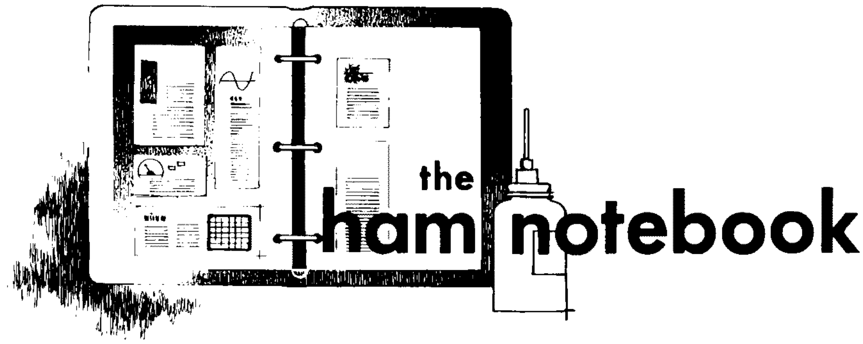
timing byte at LO memory address 001	number of clock cycles	pulse width (ms)
000	3850	1.925
001	25	0.0125
002	40	0.02
003	55	0.0275
004	70	0.035
005	85	0.0425
010	130	0.065
020	250	0.125
050	610	0.305
100	970	0.485
200	1930	0.965
300	2890	1.445
350	3490	1.745
377	3835	1.9175

For a 2-MHz microcomputer, a single clock cycle has a duration of 500 ns. The program in table 1 and associated SN7474 flip-flop provide an example of what we mean by "the substitution of hardware by software" namely a simple program and a single flip-flop replace a much more complicated hardwired programmable monostable circuit.

reference

1. *Bugbook III. Microcomputer Interfacing Experiments Using the Mark 80 Microcomputer, an 8080 System,* (E&L Instruments, Inc., Derby, Conn., 1975. Available for $14.95 from Ham Radio Books, Greenville, N.H. 03048).

ham radio



keyer modification

Many CW operators who use electronic keyers prefer a single-shaft paddle of the non-iambic variety, such as the *Vibro Key.* As they develop their speed level and go to close spacing on the paddle, many operators notice the effect of paddle bounce when using keyers with dot memory.

What happens is that after the operator strikes the paddle for a dash, the return motion of the paddle overshoots and causes the dot mechanism to close momentarily, which sets the dot memory. This, in turn, generates an unexpected (and unwanted) dot.

Careful adjustment of the paddle will minimize this effect, but it will still occur whenever the dash side is firmly hit. This is not, perhaps, a problem for the operator with a precise fist, but not all of us meet that description, and the extra bounced dot is very disturbing.

The cure is relatively simple and can be utilized with any TTL keyer. I have installed the circuit in fig. 1 in both a Data Signal 21B keyer and a keyer built around the new Curtis keyer chip, and in both cases the bounce problem was cured.

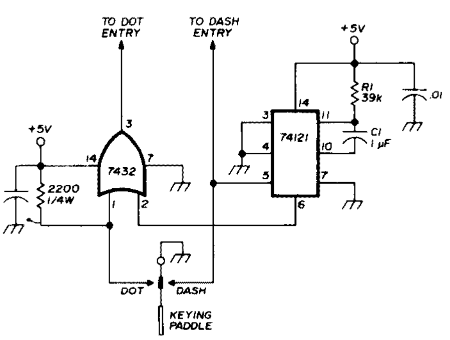


fig. 1. Simple circuit eliminates erroneous dots from being generated by paddle contact bounce (in keyers with dot memory).

The circuit uses a 74121 monostable multivibrator and a 7432 AND gate. The output of the 74121 stays low if the paddle is not in use, if dots are being sent, or if dashes are being sent. However, as soon as the dash paddle is released, the transition from the low to high state causes the 74121 to transmit a high level pulse of short duration to the AND gate. The duration of the pulse is controlled by the values of R1 and C1.

When either of the inputs to the AND gate is in the high state, the output of the gate stays high so the paddle cannot transmit a dot into the dot memory. The duration of the pulse from the 74121 is selected so that it is only long enough to block a dot caused by the dash bounce from being placed in the dot memory. The duration of the pulse is short enough that the operator cannot possibly "reverse fields" with his hand fast enough to lose a dot he intentionally sends.

In fact, with the circuit installed, the only change the operator will notice is that he no longer sends erroneous dots which are caused by the key bounce. Installation in any TTL or CMOS keyer is very simple -- the keyed lines from the paddle are fed through the circuit and connection is made to the +5 volt line. (It should be noted that some CMOS keyers use voltages other than 5 volts, in which case this circuit will not work).

The values of R1 and C1 shown in the circuit were determined experimentally, and should work fine. If you notice any blocking of intentional dots, either R1 or C1 should be reduced in value until the problem disappears.

Bob Locher, W9KNI

Collins KWM-2/KWM-2A modifications

Over the years the Collins KWM-2 and KWM-2A ssb transceiver has undergone a number of modifications, some of which were made during the period the unit was used in military service. Available through MARS libraries, and possibly the Government Printing Office, is an Air Force Technical Manual that lists over 50 modifications to the KWM-2/KWM-2A along with expanded, fold-out diagrams of the circuitry which are a great improvement (over the amateur-style instruction manual) for the bifocal crowd.

Of interest to all KWM-2/KWM-2A owners is a simple modification that consists of adding a 0.01 µF, 400-volt capacitor from the screen (pin 8) of the 6EB8 audio amplifier to ground. This eliminates an ultrasonic oscillation that caused increased noise and audio distortion in some models.

The title of this technical manual is: *TO-31R2-4-183-3.KWM-2A Transceiver.* It also covers changes to the 30L-1 and 30S-1 rf amplifiers. A second technical manual of interest to KWM-2/KWM-2A owners is TO-31R2-4-183-2 entitled, *Technical Manual (Service) KWM-2A Transceiver.* It also covers the previously mentioned amplifiers. This publication provides detailed alignment instructions for the transceiver and linear amplifiers.

William I. Orr, W6SAI

IC holders

A convenient method for storing integrated circuits in your parts cabinets is to line the bottom of the drawers with 3/8 inch (1cm) of the polyfoam packing

material used to ship electronic equipment. Push the IC leads into the soft material, keeping the ICs in neat rows and all facing the same direction. This way you can see at a glance which circuits you have in stock and keep them damage free.

Gary L. Tater, W3HUC

receiver incremental tuning for the Heath SB-102

A limited amount of receiver incremental tuning (RIT) may be obtained with the SB-102 quite easily. I own a unit with the transistorized linear master oscillator (LMO). At the rear of the LMO is a terminal marked FSK. Unless you are operating RTTY (which does not appear to be recommended in the SB-102 manual) with genuine FSK, this terminal is not used. However, it will provide up to a 1 kHz shift in frequency when directly grounded. By using the circuit shown in **fig. 2**, plus or minus 400-500 Hz shift may be obtained. As

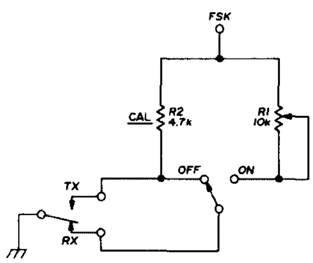


fig. 2. Receiver incremental tuning (RIT) circuit for the Heath SB-102 makes use of the built-in FSK circuit. R1 and R2 are added components.

shown, the circuit provides a useable amount of RIT which is convenient for netting ssb signals. When used in conjunction with the optional 400 Hz CW filter, it really shines.

The calibration resistor, R2, provides an essential mid-point setting and, with R1 centered, a beat note should be of equal pitch. A 10k multiturn pot could be used at R2 if extreme accuracy is desired. The dial calibration will shift approximately 400-500 Hz, but this is easily restored by the zero-set knob. The LMO shaft may be slipped slightly if you're finicky.

Paul K. Pagel, K1KXA

repairing R390 rf transformers

The rf transformers in the R390 and R390A receivers can be used for more than one band. The tuning ranges are: 0.5 to 1, 1 to 2, 2 to 4, 4 to 8, and 8 to 16 MHz. If there is a loss of sensitivity on some bands, or there is difficulty in obtaining proper alignment, here is a trouble to look for. It is more common in the R390 but may also show up in an R390A. In several first rf transformers I have found that the tuning core sticks because of lumps on the inside of the coil form. This is quite easy to check for.*

Look carefully at the rf tuning-slug racks as the mechanism tunes through its complete range on the band in question (or check them all as a precaution). The racks should move up and down smoothly. Check several times from different angles. Also check by pulling them up and down by hand while at the bottom of their range. Look for one end pushing up or causing the rack to deviate from the horizontal.

If you think there is trouble, it is easy to verify. Carefully remove the springs from each end of the slug rack and hang them out of the way under tension (use a bent paper clip). If you let go of the spring it can drop down inside the set and will be difficult to retrieve and reconnect. Lift up on the rack and it will come out quite easily. (When putting the rack back, work slowly, as it is easy to chip the edges of the coil forms when reinserting the tuning cores). When the rack has been removed, shine a light down inside the coil form and look at the side. Lumps show up immediately.

If you have this trouble it is easy to fix, but it must be done with care or the coil form will break. To remove the rf transformer, insert a Phillips screwdriver into the two little holes on the top of the transformer case and loosen the two captive screws. Then wiggle the transformer loose as you pull up. It may help to pry gently with another screwdriver.

*While the coils described are those for the Collins R390 and R390A series of receivers, there are many surplus and commercial receivers which use similar permeability-tuned mechanisms that might be susceptible to the same problem. The repair technique described here could be easily adapted to other, similar tuning mechanisms.

Don't back off on the captive screws any more than you have to. They can go past the point of releasing the transformer, come out of their mounting threads and rattle around loose inside the case. If this happens, take the top off the transformer and use a pair of needle nose pliers to hold the screws in position while you rethread them back.

To repair the coil form you have to remove the lumps from the inside. They appear to be bubbles of varnish or whatever finish was applied to the coil by the manufacturer. The first thing to do is to strengthen the coil form. To do this, spread several layers of *Elmer's Glue-all* on the outside of the form. Be sure each layer has plenty of time to dry; leave it overnight. This will give added strength to the form and coil and help keep either from breaking.

Next, go to work with your box of electric drill bits. Start with 13/64 inch (5mm). Gently insert it into the coil from and twist it by hand to begin removing the crud. When that cuts through, use a 7/32 inch (5.5mm) bit and do the same thing. Finish up with a 1/4 inch (6.5mm) bit. This will take most of it off.

Now make a tube of emery paper (fine sandpaper might work) long enough to reach to the bottom of the coil form and still leave a hand hold. Insert that into the coil form. Take a drill bit thin enough to slip easily inside emery paper but thick enough to give it support. Twist the emery paper around inside the coil form, moving it up and down at the same time. This will smooth off the inside again.

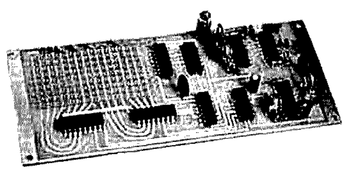
The thing to watch out for here is that you don't chip the top edge of the coil form. If it is chipped or looks about ready to go, it can be strengthened with a thin strip of typewriter paper, spread with *Elmer's Glue-all*, wrapped a few turns around the outside of the coil form top.

Every so often, remove the drill and emery paper and try the tuning core back inside. It should move up and down the entire length of the coil form without binding. When you have completed the operation, clean out the emery and coil form dust by blowing or using a pipe cleaner. Before you put the slug rack back in the set, give the *inside* of the coil forms and the tuning cores a squirt of silicon spray.

Alexander MacLean, WA2SUT

NEW products

programmable cw identification kit



With the new CW ID kit offered by VHF Engineering, you can build a complete identifier for commercial or amateur repeaters in about one evening. The CW ID kit uses high-grade components and comes complete with a drilled epoxy-glass circuit board and programming diodes.

Sufficient diodes are included to allow you to program virtually all repeater calls. To program, all you do is solder the appropriate diodes directly to a matrix on the PC board furnished. The diodes are mounted on the board in a straight-line fashion: three diodes for a dash, one for a dot, and none for a space. Programmed calls can be changed easily by rearranging the diodes. You can program the board for either CW or RTTY, which means added flexibility.

The CW ID kit is available for $39.95 plus postage; wired and tested it's $49.95 plus postage from the manufacturer. Drop a note to VHF Engineering, 320 Water Street, P.O. Box 1921, Binghamton, New York 13902 for more information, or use *check-off* on page 126.

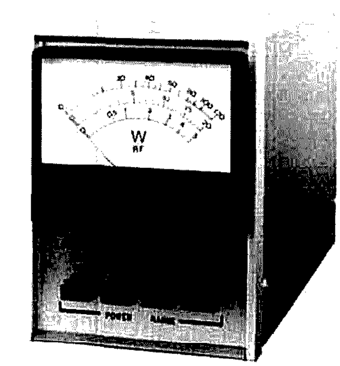
vhf wideband preamplifier

Twin-output preamplifiers are something new offered by Spectrum International for the vhf buff. Designated MMa50, MMa144, and MMa220 (for the 50-, 144- and 220-MHz bands), these preamps feature two untuned stages with twin outputs for feeding two independent receivers. The preamps are built on a glass-epoxy G10 PC board, which is mounted in a standard die-cast aluminum box. Gain and noise figure are quite respectable as shown in the following table:

	MMa50	MMa144	MMa220
freq range, MHz	50-54	144-148	220-225
nominal gain, dB	20	16	15
noise figure, dB	2.5	2.8	3.4

Power requirements are 12 volts dc at 20 milliamperes; size is 1¼ x 2½ x 4½ inches (32x64x114mm). The specifications apply to a 50-ohm input-output system. The MMa50 and MMa144 sell for $29.95 each; the MMa220 for $34.95. Add $1.00 shipping charge for each unit. Write Spectrum International, P.O Box 1084, Concord, Massachusetts 01742 for more information, or use *check-off* on page 126.

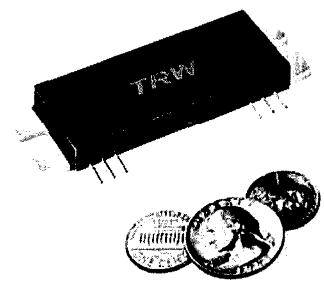
rf wattmeter



The trend today in radio transmitters is to have a wattmeter in the transmission line feeding the antenna. The model LPM-880, a new product by Leader Instruments Corporation, is a direct-reading wattmeter that measures radio-frequency power in the 0.5-120 watt range. Power range is selectable by front-panel pushbuttons. Also included is a dummy load for off-the-air power measurements. You can use the LPM-800 for measuring power loss in low-pass filters and coax cables as well as transmitter output. It's supplied with a sturdy tilt stand for easy reading. Input impedance is 50 ohms.

The LPM-880 is priced at $149.95. If you'd like more information contact Pat Redko, Leader Instruments Corporation, 151 Dupont Street, Plainview, Long Island, New York 11903 or use *check-off* on page 126.

power modules for mobile transmitters



Two new vhf rf power modules designed for mobile or marine transmitter applications are now available from TRW Semiconductors. The modules, designated MV20 and MV30, provide in excess of 20 watts and 30 watts output power respectively across the 140-175 MHz band. The modules operate from standard 12-volt automotive supplies and withstand infinite vswr at any angle, with 2 dB overdrive and 16 volts dc applied. The modules also feature 50-ohm input and output impedances, more than 20 dB gain, and are stable when operating into load vswrs as high as 5:1.

When compared to discrete component designs, these modules offer significant savings in size as well as cost of design, production, and repair. Small quantity pricing is $39.50 for the MV20 and $41.50 for the MV30. For further information, contact Sales Manager, Mobile Products, TRW RF Semiconductors, 14520 Aviation Blvd., Lawndale, California 90260 or use *check-off* on page 126.

loudspeaker for voice communications

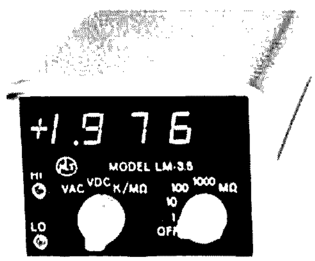
The Kriket[R] model KC-55 speaker, new from Acoustic Fiber Sound Systems Incorporated, is designed to reproduce the human voice with maximum intelligibility. It has a frequency

More Details? CHECK—OFF Page 126

response between 80-10,000 Hz, essential for voice communications. The secret is in the Kriket[R] 5-inch-diameter (127mm) permanent-magnet speaker and an exclusive AFS Working Wall[R] enclosure, which controls sound by eliminating distortion. A snap-lock mounting bracket permits adjusting the speaker in any desired direction. The KC-55 speaker is packaged for base-station use, but you can adapt it for mobile use simply by removing the base for easy mounting inside a vehicle.

The KC-55 handles 7 watts rms of audio, has an input impedance of 8 ohms, and is furnished with a 6-foot (1.8m) connector cord and a standard miniplug. More information is available from acoustic Fiber Sound Systems, Incorporated, 2831 North Webster Avenue, Indianapolis, Indiana 46219, or use *check-off* on page 126.

digital multimeter



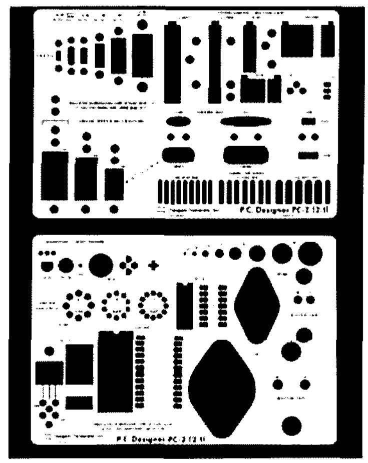
Non Linear Systems of Del Mar, California, announces a new addition to their Volksmeter family. It's the LM-3.5 Volksmeter Plus, a 3½-digit multimeter that fits into the palm of your hand. The 3½-digit feature means that the instrument reads out three digits plus 100 percent overrange. It's a true multi-function, multirange meter, rugged enough for field use yet useful for production or hobby work. Rechargeable nicad batteries and a 115-volt charger are standard equipment.

The LM-3.5 has four ranges for dc and ac volts, to 1000 volts dc or 1000 volts peak ac, with 1-millivolt resolution on the 2-volt scale. The resistance scale has one-ohm resolution and five ranges, from 2000 ohms to 20 megohms full scale. Ac and dc current can be measured in three ranges using shunts furnished. Automatic polarity is featured. Input impedance is 10 megohms on all voltage ranges. A large light-emitting diode display (0.3 inch or

7.6mm) and small package size of 1.9 x 2.7 x 4 inches (48x69x102mm) is a good example of what can be done with large-scale integration technology.

The LM-3.5 Volksmeter Plus retails for $147.00, including input leads, rechargeable nicads, battery charger and current shunts. Optional accessories, such as carrying case, high-voltage probe, desk stand, panel-mount flange, and universal test-lead set are also available. For information on these accessories as well as other data, write Non Linear Systems, Incorporated, P.O. Box N, Del Mar, California 92014, or use *check-off* on page 126.

pc design template



The new *PC Designer* template set has been updated to include additional, frequently used component packages and mounting patterns for printed-circuit layouts and assembly drawings. All patterns conform to guidelines established by Mil-Std 275C and The Institute of Printed Circuits bulletin CM-770. Component mounting patterns are on grid centers which enable the designer to design for automatic insertion assembly equipment.

Fixed and variable resistors, axial and radial lead capacitors, and several semiconductor packages are included. Template use reduces circuit-board design time by eliminating constant referral to manuals and data sheets for package dimensions. The template sets are available from stock in actual, twice, and four times size layout ratios and are priced from $12.00 to $20.00 per set. Write Tangent Template, Inc., Post Office Box 20704, San Diego, California or use *check-off* on page 126.

More Details? CHECK—OFF Page 126

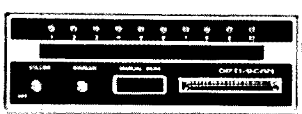
multipurpose cw-operating aid

From SEA International comes the RST 599, an instrument that combines four features in one to enhance CW operation. Packaged in an attractive cabinet, the RST 599 includes a universal keying monitor, code-practice oscillator, CW filter, and a feature called the 599 function.

The RST 599 can be used with any key, keyer, transmitter or transceiver. The CW filter has a nominal center frequency of 850 Hz, with a nominal 3-dB bandwidth. The 599 function extracts and reconstructs the selected signal and, according to the manufacturer, "makes it noise free and RST 599 every time." Inputs are *audio in,* for connection to any receiver output line, and *key in,* for connection to any key or keyer. Outputs are *speaker out,* which drives any speaker of 4 ohms or more; *headphones,* for phones of 4 ohms or more and *transmitter,* which connects your key to any transmitter or receiver. When the 599 function is switched off, receiver output is connected directly to your speaker.

The RST 599 is priced in the $90.00 range and is fully warranteed for one year against material or manufacturing defects. If you'd like more information, write SEA International, P.O. Box 32, Milpitas, California 95035, or use *check-off* on page 126.

programmable scanning receiver



The Opti/Scan, a 10-channel scanning monitor receiver by SBE (Linear Systems, Incorporated), offers some unusual features that will appeal to the vhf enthusiast. Frequencies are digitally synthesized, which means you can forget about buying crystals to obtain desired coverage. You program the receiver yourself to scan channels of interest. Programming is easy. You simply refer to a code list supplied for desired scanning frequencies, program a

plastic card according to instructions, insert the card into a slot on the receiver front panel, and internal circuits do the rest. No need to worry about programming errors – you can see what you've programmed and can check for accuracy. As many cards as desired may be programmed – up to 16,000 frequencies may be selected. Program cards are only $2.25 each.

Any ten channels are available between 30-50, 150-170, 450-470, and 490-510 MHz. Two-meter-band coverage (140-160 MHz) is available on special order. The Opti/Scan receiver is furnished with antennas that cover the vhf-uhf utility bands mentioned above. Receiver sensitivity is 0.5 microvolt for 12 dB SINAI. Dimensions are 2½ inches high, 7-3/4 inches wide, and 10 inches deep (6.4x19.7x25.4cm). Operating voltages are 13.8 Vdc or 115 Vac 60 Hz. The list price is $369.95. For further information contact Linear Systems, Incorporated, 220 Airport Boulevard, Watsonville, California 95075, or use *check-off* on page 126.

1976 Allied catalog

Now available from Allied Electronics is the 1976 *Engineering Manual and Purchasing Guide.* This up-to-date manual is a must for the service bench or engineering library. Engineers, radio amateurs, technicians, and hobbyists will appreciate the complete specifications, illustrations and information which describe each product.

This newest edition of the well-known Allied Guide offers 228 pages of high-quality electronic parts and equipment from Allied and other leading manufacturers. You can choose from a wide variety of new products, in addition to traditional items which have set in the industry's standards. The guide contains cable, solid-state devices, test equipment, connectors, relays, tools, capacitors, and countless other electronic parts for virtually any application. Allied offers bulk pricing for quantity buyers, and nationwide warehouses assure prompt delivery of these often hard-to-get items.

A $5 value for only $1 to help cover postage and handling. Contact Allied Electronics, Dept. 76, 401 East 8th Street, Fort Worth, Texas 76102 or use *check-off* on page 126.

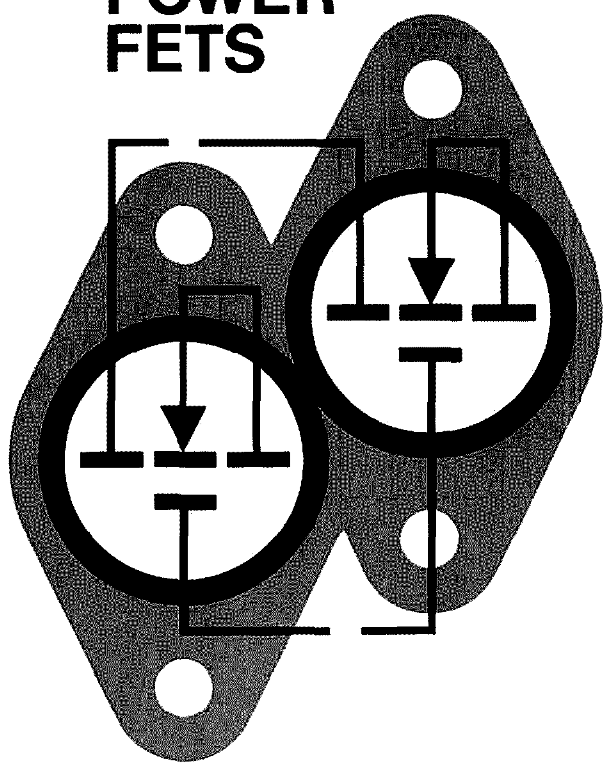
More Details? CHECK–OFF Page 126

one dollar

ham radio magazine

SEPTEMBER 1976

TWO·METER TRANSVERTER using POWER FETS



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contents

a second look
by Jim Fisk

If attendance at the microprocessor seminars at the various hamfests around the country is any indication, amateur interest in these versatile machines is growing by leaps and bounds. Although a good deal of this increased interest is due to the drastic price reductions of the past few months, improved support from the manufacturers in terms of hardware and software have been a contributing factor. Whereas last year's computer hobbyist spent the majority of his time developing and building hardware, today the trend is toward the software, or programming, side of computer design. And, as many hobbyists are discovering, software is often more challenging – and rewarding – than wiring up a board full of logic chips.

Another trend that promises to increase the popularity of the home computer is that of designing the basic machine to accomodate future developments in microprocessor technology. Although the direct interface logic is different for each processor chip, with careful circuit design it's possible to retain all of the expensive memory boards of the basic system, as well as most of the input/output circuitry. This takes a good deal more planning, and an understanding of LSI IC design, but the computer manufacturer who is successful will dominate the future market. As it stands now, most computer systems are oriented around a single microprocessor chip. This is unfortunate because it means that each new, improved processor that reaches the marketplace requires a whole new design, an entirely new machine – input/output logic, memory *and* power supplies. At least one company, The Digital Group, Inc., has recognized this basic deficiency and has come up with a computer system that will accomodate a wide range of microprocessor chips. With this design concept you don't have to buy a whole new system each time an advanced processor hits the market; you simply buy a new CPU card (at considerable cost savings) and use your existing memory and input/output logic.

As with all new developments, the microprocessor field has been in a continual state of flux since the first device was announced several years ago, and I see no reason for it to settle down in the near future. If you made a survey of computer hobbyists right now, you'd probably find that most of them use an Intel 8080 or 8080A in their system, with the Motorola 6800 running a distant second. This is an important consideration if you're thinking about buying a computer for home use because it means that there are a large number of programs available which you can put to work as soon as you get your machine. Once you've had some "hands on" experience with existing programs, the task of writing your own programs will look much less formidable. That's when the real fun of owning your own computer begins.

It's important to keep in mind that just because the 8080 is king of the hill right now doesn't mean it's always going to be that way. The new Zilog Z80, for example, operates much like the 8080, but is faster and has a more powerful instruction set; waiting in the wings are yet more powerful processor ICs which will reach the market in coming months. However, if you choose a system that has good software support, and is designed to accomodate future developments in microprocessor technology, you'll have a computer that can be easily updated and expanded as soon as devices become available.

Many amateurs have been watching the home computer market with caution, waiting for the product to mature to the point where they can buy the most computing power for the least cost. Although we'll almost certainly see some further price reductions in the months and years ahead, today you can put together a very good system for about the same cost as an ssb transceiver. If you choose wisely, and spend your money carefully, you can have a system *now* that will meet practically all of your future computer needs.

Jim Fisk, W1DTY
editor-in-chief

presstop

LONG-AWAITED CB EXPANSION was announced July 27th with 17 new channels added to the existing 11-meter CB band. The expansion to 40 channels won't become effective until January 1 in order to give manufacturers time to develop, and have type-accepted, the new radios. One of the 17 new channels falls between channels 22 and 23 — the remaining 16 are from 23 up.

Tough New Technical Specifications on all new 40-channel CB gear are also a part of the Report and Order on Docket 20120. All transmitter harmonic and spurious radiation must be down 60 dB from the carrier instead of the present 49 dB, and total receiver radiation at the antenna terminals can't exceed two nanowatts. In addition, receiver chassis radiation must be under five microvolts at three meters. As tough as these new specs seem, however, they are likely to get tougher.

TYPE ACCEPTANCE OF AMATEUR RADIO GEAR is still very much under consideration by the FCC. It's not only a result of the misuse of Amateur transceivers by the "HF" groups and the proliferation of "3-30 MHz" broadband linear amplifiers, but is also due to some serious interference problems traced to inaccurately "spec'ed" commercially-made Amateur gear that was being used by Amateurs. Some action along these lines is not far off — perhaps by late Fall.

NOVICE PHONE PRIVILEGES are to be proposed to the FCC by the ARRL. The League Board of Directors voted in Denver to petition the Commission to immediately add 145-145.5 and 222-225 MHz CW, AM and FM with 50 watts DC input to the Novice frequency bands.

EXPERIMENTAL NOVICE LICENSING PROGRAM became an official reality with an FCC Public Notice issued July 21. Only a limited number of organizations will be selected to participate in the experimental phase of the program, which is designed to be a "controlled experiment" leading to improvements in the Novice license examination system.

Organizations Interested In Participating in the FCC's new Novice license examination program should submit a detailed proposal to John Johnston, Chief, Amateur and Citizens Division, FCC, Washington D.C. 20554 for review.

Minimum Course Coverage for the experimental program must include all material in the Commission's Element Two Novice Study Guide but may be taken from any source. Instructors must be experienced and knowledgeable, hold a General or higher class Amateur license and be 21 or older. Course graduates will be required to pass a five wpm code test and a written examination of at least 20 questions covering the nine Element Two Study Guide categories — and some students from each class will still have to take a standard Commission-graded Novice examination.

AMATEURS APPLYING to the FCC for license renewal, modification, or other action should always try to pay by personal check. The returned cancelled check provides assurance that your application made it through the mail, and may even prove you've paid if your paperwork later gets lost in the system.

"Straight Forward" License Processing is now down to 7-8 weeks for most applicants, according to late reader reports. Anyone who has been waiting longer than 12 weeks for an Amateur license should probably call FCC at (202) 632-7175 for help.

1X2 CALLSIGN REQUESTS INCREASED SHARPLY after a very slow beginning. Since 1x2 assignments are hand processed, some delays seem likely — even some pre-July 1 requests for unspecified 1x2 calls will probably be delayed by the influx.

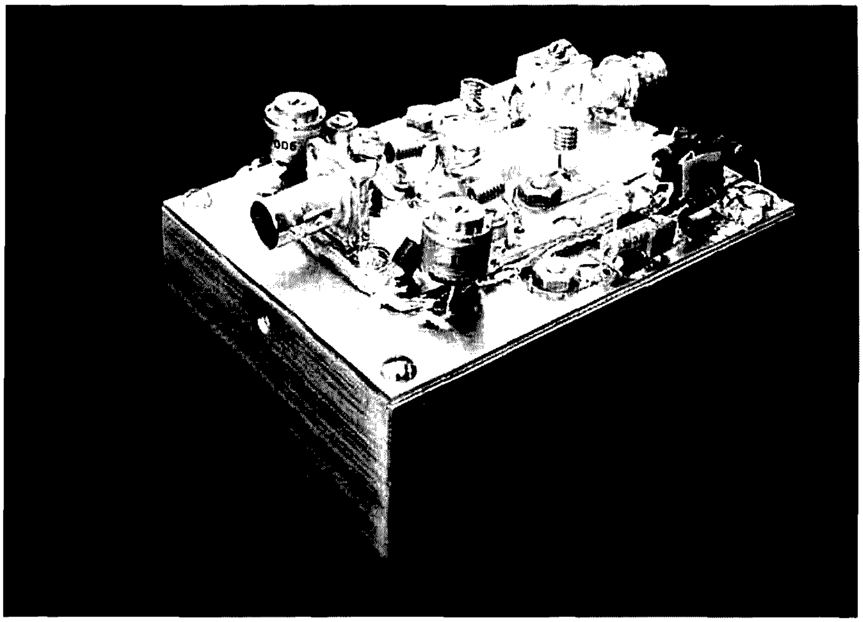
WD8 CALLSIGNS have been issued in quantity and should be showing up on the bands momentarily. A look in the latest Callbook shows WD4, WD9 and WDØ shouldn't be far behind. For bicentennial purposes WD call holders should use "AE" for a prefix.

WT-PREFIXED CALLS have been issued to a few Amateur license applicants who've been the victims of an extreme foulup in the FCC's paperwork mill. The WT (for temporary) calls are good only until the computer issues the new Amateur's permanent call, so they should be very good catches for prefix hunters.

OSCAR 8 DELAY now looks like it might extend through 1980 for want of a timely launch vehicle. An interim satellite to supplement OSCAR 6 and 7 is being considered to take advantage of a NOAA launch opportunity in late 1977. Experienced volunteers to work on both hardware and software for the project will be needed.

Coast Guard Cutter planning an extended Arctic tour wants to use OSCARs for crew communications because of HF propagation problems and its being too far north to use synchronous satellites over the equator. AMSAT is looking for volunteers to make coverage plots to determine practicality — W3GEY has details.

WWV'S REMAINING PROPAGATION REPORT will be discontinued October 1 unless a tentative decision to disband the Telecommunications Services Center, which supplies the 14 minutes after the hour report, can be reversed. Write Dr. Douglass Crombie, Director ITS/OT, U.S. Department of Commerce, Boulder, CO 80302 or call him at (303) 499-1000, ext. 4215 — a carbon of your letter to the Honorable T.E. Wirth, U.S. House of Representatives, Washington, D.C.



two-meter transverter

using power fets

How does 10 watts
PEP output for
1 milliwatt input sound?
Here's a complete transverter
using a new power fet
that gets it all together

This article describes a two-meter transverter which, with approximately 1 milliwatt input, will produce a nominal 10-watt PEP output with all distortion products 28 dB down from one tone of a two equal-tone test signal. The receiving converter, which uses a conventional dual-gate mosfet rf amplifier, compares favorably with a commercial receiver of recent design. These units are designed for use with a 28-MHz transceiver (in my case the Kenwood Twins).

design considerations

I used a modular approach in the design of this transverter. Advantages of this approach are given in an excellent article by Joe Reisert.[1] One of the advantages of the modular approach becomes apparent during initial alignment, because three different output levels are readily available: from the mixer module, a nominal 100 milliwatts; from the first linear amplifier, a nominal 2 watts; and from the final linear amplifier a nominal 5 or 10 watts. The intermodulation specification for the two lower-level modules is a nominal 35 dB down. All modules are designed to work into a 50-ohm load. A block diagram is presented in fig. 1.

By Larry Leighton, WB6BPI, Siliconix, Inc., 2201 Laurelwood Road, Santa Clara, California 95054

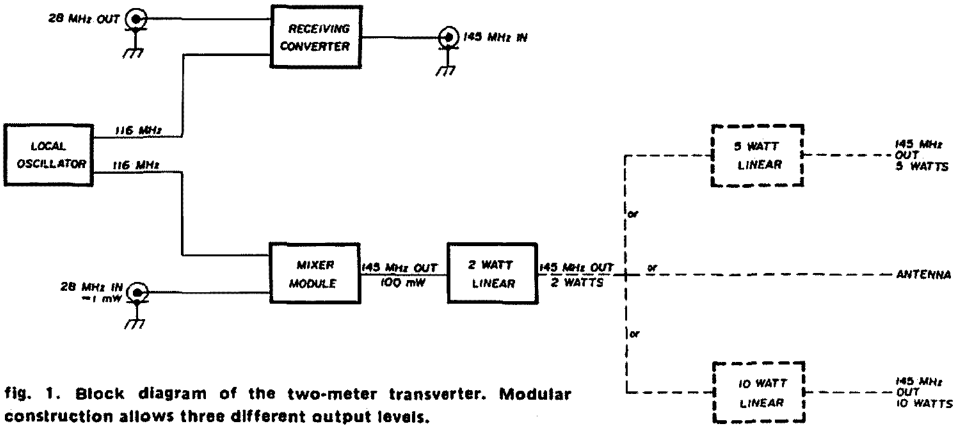


fig. 1. Block diagram of the two-meter transverter. Modular construction allows three different output levels.

Oscillator. This circuit (fig. 2) is described generally in reference 2. It is designed for use with a fifth-overtone 116-MHz crystal. In the past I've experienced difficulty with overtone oscillator circuits but this oscillator circuit is an exception. I've encountered no problems with spurious outputs or hard starting; and with the dual-gate buffer, the output of this module is relatively clean. The inductor in parallel with the crystal is necessary to prevent spurious oscillations. The oscillator module provides two mutually isolated 50-ohm outputs at 116 MHz. Nominal output level is 1 volt rms per output port. Using this approach and a little care, the transmitting converter can be used to tune the receiving converter provided you have a separate transmitter and receiver.

The oscillator output network is a lumped-constant equivalent of a Wilkinson n-way combiner, fig. 3. The constants for this circuit can be determined from the following equations.

$$C1 = \frac{1}{2\pi f_o R_o} \quad (1)$$

$$C2 = \frac{C1}{2} \quad (2)$$

$$L1 = L2 = \frac{R_o}{2\pi f_o} \quad (3)$$

$$R1 = 2R_o \quad (4)$$

where:

f_o = frequency (Hz)

$R_o = R_{gen} = R_{load}$ (ohms)

C = capacitance (F)

L = inductance (H)

For example, using f_o = 116 MHz and R_o = 50 ohms,

$$C1 = \frac{1}{(2\pi)(116 \times 10^6)(50)} = 27\ pF$$

$$C2 = 27/2 = 13.5\ pF$$

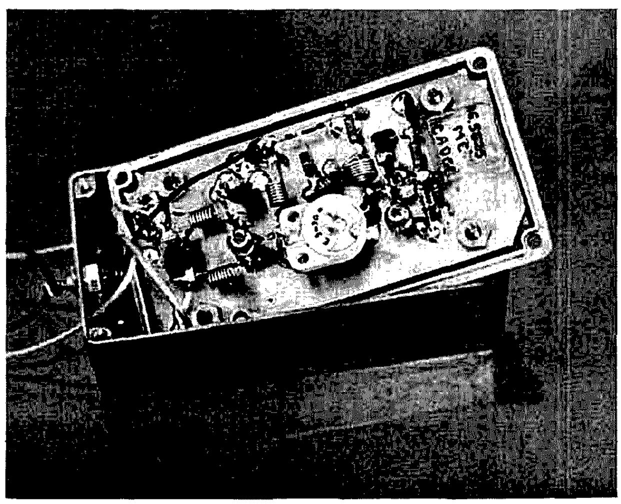
$$L1, L2 = \frac{50}{(2\pi)(116 \times 10^6)} = 69\ nH$$

$$R1 = 2 \times 50 = 100\ ohms$$

Mixer. The mixer circuit (fig. 4) has many advantages over conventional doubly-balanced mixers including low component count, and no requirement for balanced transformers. The mixer portions provides approximately 1 dB conversion gain and has a nominal 50-ohm input impedance at both input ports. The 1k variable resistor is used to minimize the 116-MHz local oscillator signal at the mixer output. While minimizing the 116-MHz local-oscillator signal, the fifth harmonic of 28 MHz is also minimized.

The input signals are cancelled at the mixer output in a manner similar to that of a push-push doubler. The mixer output contains the beat signal plus even-order

Local-oscillator module. Point-to-point wiring is used, with component leads serving as tie points. The trimmer capacitor is adjusted for maximum negative voltage at test point 1 (see fig. 2).



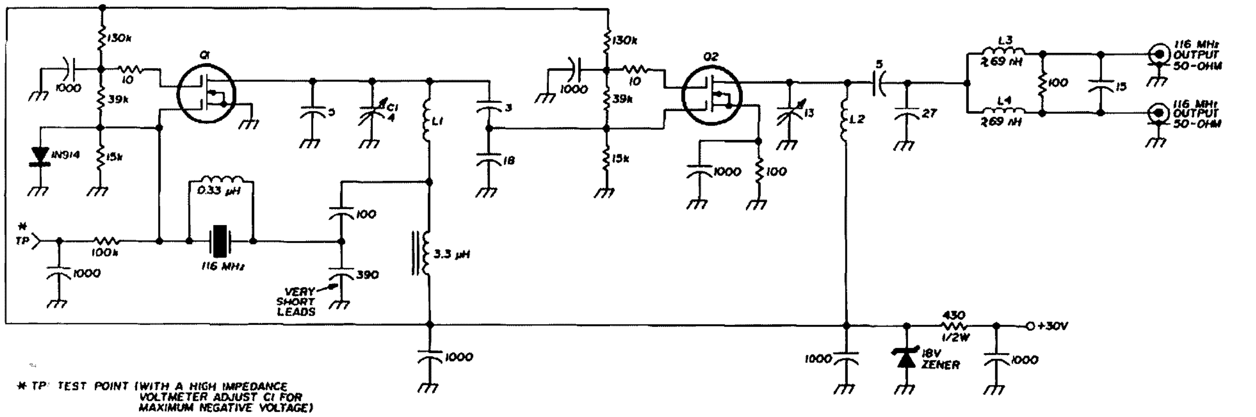
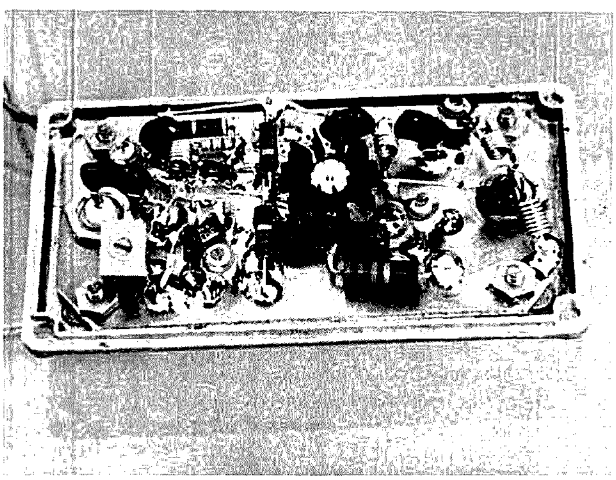


fig. 2. Local-oscillator module provides two mutually independent outputs at 116 MHz. L1 is 5 turns, ¼" (6.5mm) ID; L2 is 7 turns, 1/8" (3mm) ID; L3, L4 are 4½ turns, 1/8" (3mm) ID. All are close wound using no. 20 AWG (0.8mm) enamelled copper wire. Q1, Q2 are Fairchild FT0601 or RCA 40673.

products of the input signals. The even-order products are attenuated by the output-network Q. The mixer circuit is very flexible. By removing the input matching networks and shunting the fet gates to ground with 10-k resistors, the choice of input frequencies can be changed easily. Only a small loss in conversion gain will be noted, and the input impedances will be something like 200 ohms. If you wish, you can design matching networks to make the input impedance look like 50 ohms again. For a more detailed treatment of this circuit, see reference 3.

The buffer transistor in this circuit draws 40 mA nominal current and dissipates approximately 0.8 watt. Because the transistor is not a high-power device, a clip-on heatsink should be used and some ventilation provided. The filter at the output of this stage is designed to attenuate the local oscillator second harmonic and could be eliminated if a bandpass filter is used at the antenna.

Mixer module. Input connector is at left, with balanced fet stage slightly to the right. The 2N3866 power stage is at the center, mounted in a clip-on heatsink. Output connector is at right.



Amplifiers. The linear amplifiers (figs. 5 and 6) provide three different output levels. One amplifier is designed to be driven directly from the mixer module. With 1 milliwatt input to the mixer module, this amplifier delivers a nominal 2 watts PEP output.

In this configuration, and with some slight changes in biasing networks; operation can be achieved with a

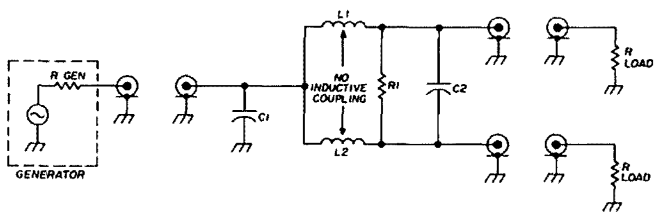


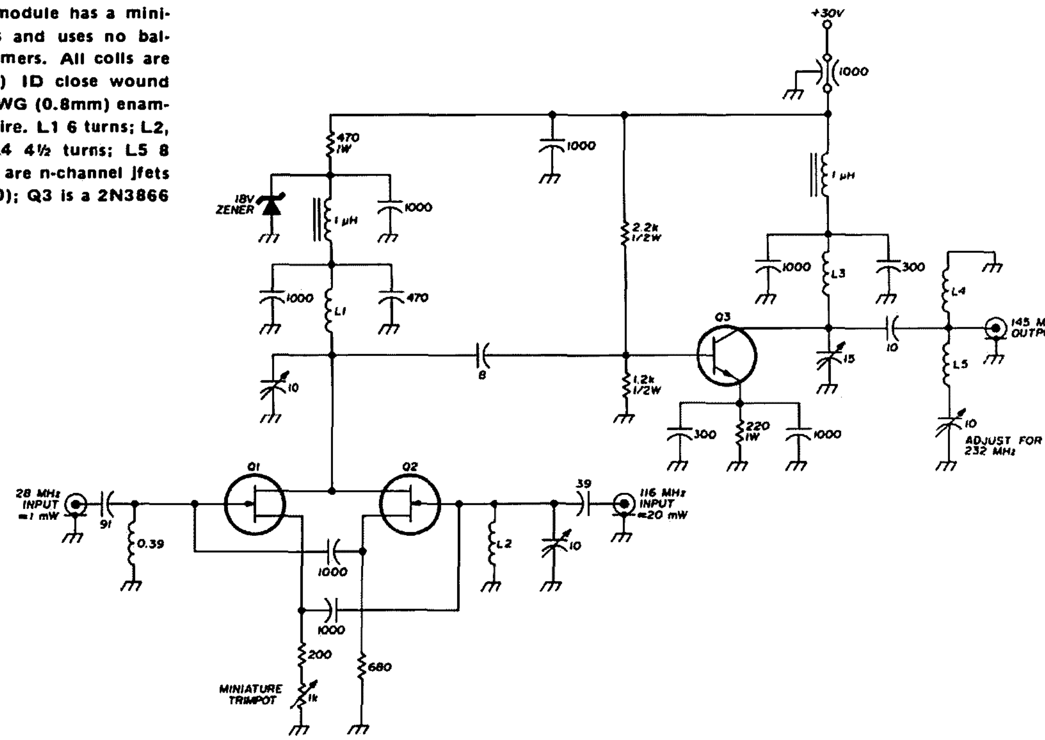
fig. 3. Wilkinson n-way divider or summer used in local-oscillator output as a lumped-constant equivalent circuit. Examples for determining values are given in the text.

24-volt power supply and only a slight decrease in output power. In both power fet circuits, bias resistors are adjusted to set the fet quiescent drain current to 150 mA. After the quiescent drain current has been set, the gate voltage can be measured, and the variable resistors can be replaced with fixed resistors.

The amplifier in fig. 5 will deliver a nominal 5 watts PEP output with a 30-volt supply. With two of these amplifiers in series, mixer drive must be reduced to keep the final amplifier in the linear region.

The amplifier in fig. 6 will deliver a nominal 10 watts PEP output and also requires a decrease in drive at the mixer. Tuning this amplifier is a bit more complicated and requires back-and-forth adjustments between the two input networks for maximum undistorted output. Both amplifiers have a nominal 12.5 dB of gain and are capable of more output than that specified; but in the interest of good operating techniques, output power should not exceed that which is specified.

fig. 4. Mixer module has a minimum of parts and uses no balanced transformers. All coils are 1/8 in. (3mm) ID close wound with no. 20 AWG (0.8mm) enamelled copper wire. L1 6 turns; L2, L3 4 turns; L4 4½ turns; L5 8 turns. Q1, Q2 are n-channel jfets (Siliconix J310); Q3 is a 2N3866 transistor.



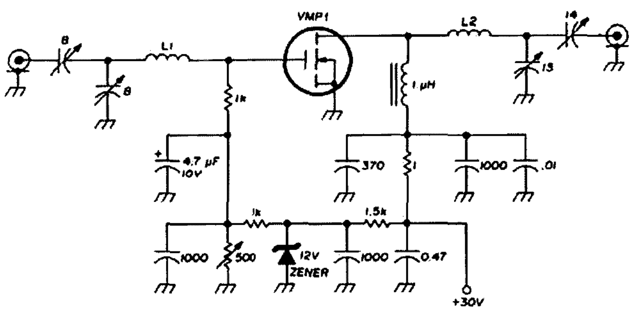


fig. 5. Two- or five-watt linear amplifier uses a Siliconix VMP1 fet. Although designed for switching service, this device works well as a linear rf amplifier. L1 is 8 turns 1/8" (3mm); L2 is 5 turns, (3mm) ID close-wound with no. 20 AWG (0.8mm) enamelled copper wire.

the power transistor

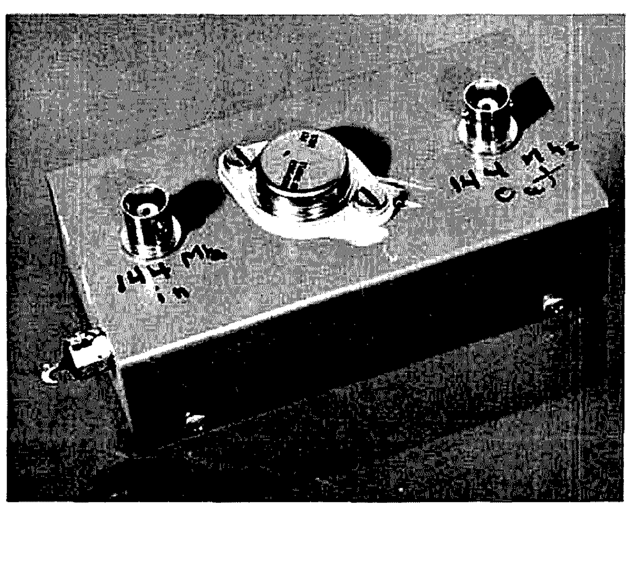
Recently Siliconix announced the availability of a high-current, high-speed switching transistor known as the VMP1 *Mospower* FET.* Although no rf specifications are available, the VMP1, designed to switch 1 ampere in 5 nanoseconds, works very well as a linear amplifier.

The VMP1 has some very desirable features compared with bipolar transistors in this application: a) no thermal runaway, b) no secondary breakdown, and c) input and output impedances are relatively high.

As with all fets, gain decreases slightly with increased temperature. Thermal runaway is no problem with the VMP1 which outweighs the disadvantage of the slight gain decrease. All three amplifiers were run for one-half hour continuously at full rated output with less than 0.5 dB decrease in output power with input level held

*For ordering information on the Siliconix VMP 1 Mospower Fet, write to Ed Oxner, Siliconix, Incorporated, 2201 Laurelwood Road, Santa Clara, California 95054.

Five-watt linear amplifier module using a Siliconix VMP1 Mospower fet. Construction of the two-watt linear amplifier is similar.



constant. At the end of these tests, the fets were only warm to the touch.

After tuning and optimizing the matching networks, I disassembled them and measured the component values. With these values and the appropriate formulas, I determined that, for best performance at 7 watts PEP output, the source impedance is 12 ohms in series with an inductive 25 ohms and remains relatively constant for

tor without introducing too much additional capacitance. In this case, I used a beryllium-oxide heatsink insulator 0.062 inch (1.6mm) thick. In this configuration, the fet shunt output capacitance was calculated at approximately 20 pF, and that of the heatsink insulator 26 pF, for a total shunt capacitance of 46 pF. At 145 MHz this presents no problem. The VMP1 can be used with circuit Qs as low as 2 using the above technique.

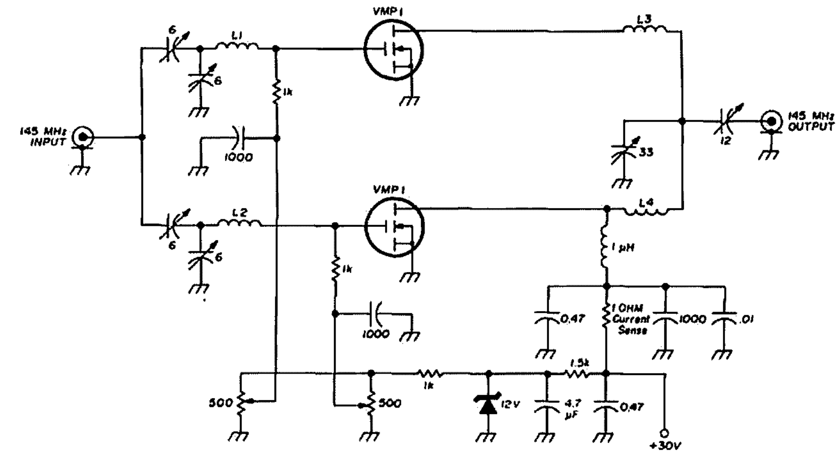
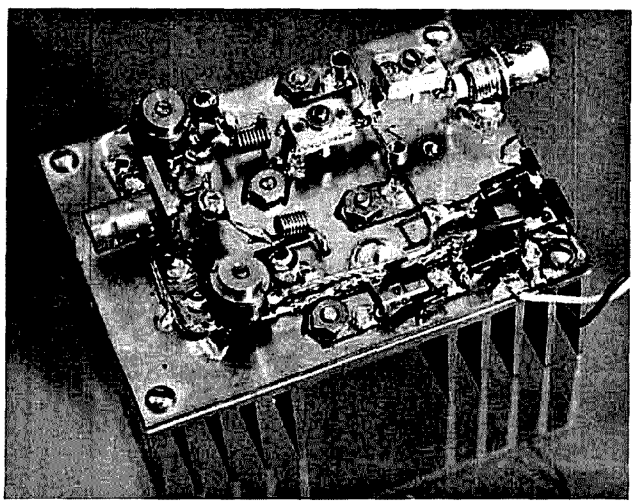


fig. 6. Ten-watt linear amplifier uses two Siliconix VMP1 for a nominal 10 watts PEP output. These amplifiers have excellent stability and minimal gain decrease with increased temperature. L1, L2 are 8 turns; L3, L4 are 5 turns, all 1/8" (3mm) ID closewound with no. 20 AWG (0.8mm) enamelled copper wire.

smaller drive levels. The load impedance for the same conditions is 12 ohms in series with 14 ohms inductive reactance.

Note that the largest contribution to output capacitance is the result of mounting the fet. Because the drain of this fet is connected to the transistor case, it's necessary to develop some means to heatsink the transis-

Ten-watt linear amplifier module uses two Siliconix VMP1 Mospower fets on a finned heatsink. In this photograph the input is at left, output at right. Bias adjust pots are at left.



The two higher power output stages have efficiencies of 30 to 40 percent and require adequate heatsinks. For the 10-watt unit, I used a 4 x 3 x 1-1/2 inch (10x7.6x4cm) fin-style heatsink with a beryllium-oxide heatsink insulator 0.062 inch (1.6mm) thick. If you don't use this style heatsink insulator, you must change the matching networks to compensate. I strongly recommend this style of insulator. For the 2-watt unit I used the recommended heatsink insulator but used the module chassis for the heatsink. (Care must be used when working with beryllium-oxide insulators. Pulverized particles are poisonous if breathed).

receiver

The receiving converter (fig. 7) has a lightly coupled interstage bandpass filter and is rather narrowband. For *full* coverage of the 2-meter band, the inductors in the interstage bandpass filter should be tightly coupled. This receiving converter will oscillate unless terminated properly at the input, which means that either the voltage to the dual-gate mosfet must be removed while transmitting, or the receiving converter input must be terminated while transmitting.

construction hints

As with any vhf project, lead lengths should be as short as possible. Circuit layout is important; if you use the schematics and photos as guides for component placement, you shouldn't have any difficulties. The values for variable capacitors are nominal. That is, these

values must be within the range of the capacitors. This allows some flexibility during parts procurement. I recommend high-quality capacitors in the linear amplifier output networks.

Initial adjustments should be made with reduced input drive. Then the input drive should be increased slowly until final adjustments can be made. In any case, should show a slight dip when the output shunt capacitor is tuned to resonance.

Any of these configurations will allow full coverage of the 2-meter band. The output in the 10-watt configuration was measured on a Hewlett-Packard spectrum analyzer. Three frequencies were visible on the display. The local oscillator output was 35 dB down from the

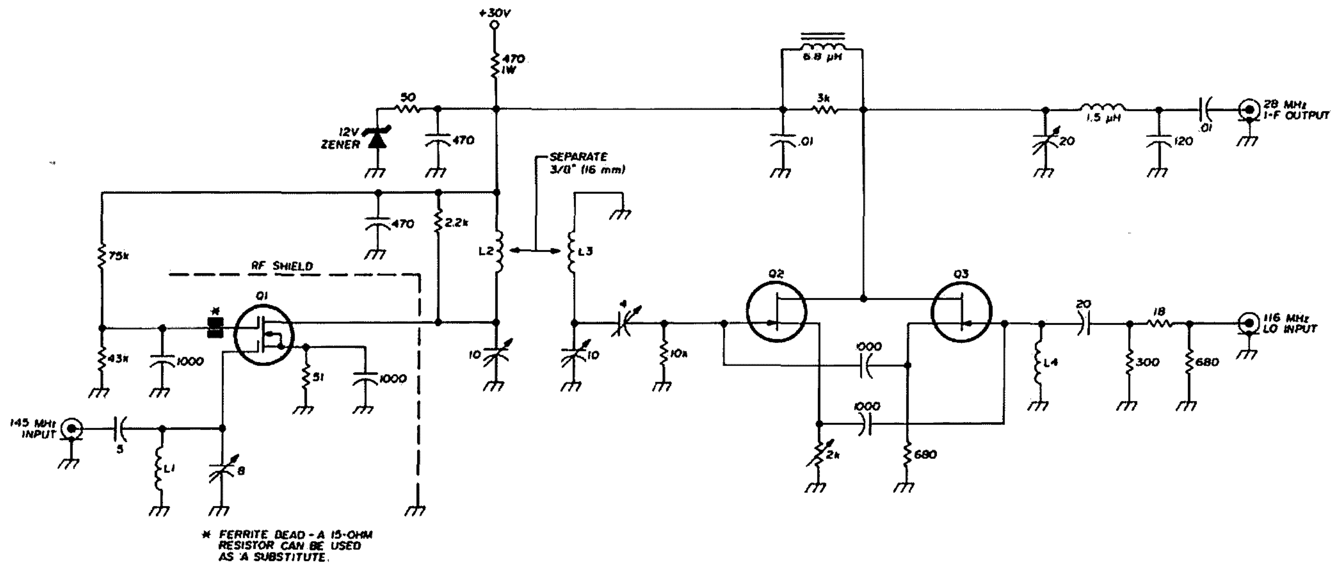
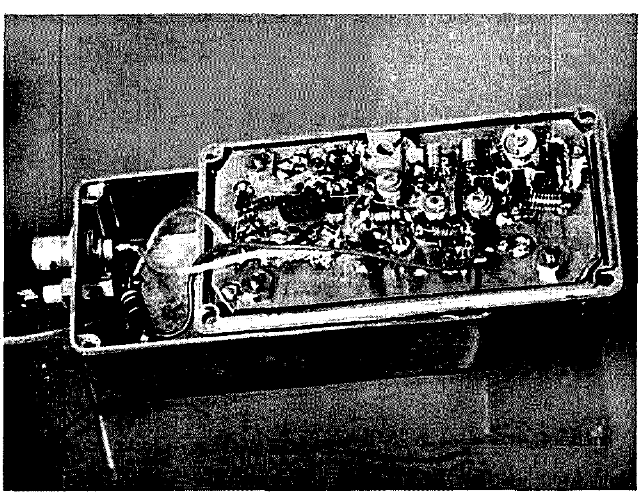


fig. 7. Receiving converter is relatively narrowband. For full coverage of the two-meter band, L2 and L3 should be tightly coupled. L1, L4 are 6 turns; L2, L3 are 8 turns. All are 1/8" (3mm) ID close-wound with no. 20 AWG (0.8mm) enamelled copper wire. Q1 is a Fairchild FT0601 or RCA 40673; Q2, Q3 are n-channel jfets (Siliconix E300, J310, or 2N5485, 2N5486.

with one exception, all variable capacitors are tuned for maximum output. If you have the proper test equipment, the two output capacitors can be adjusted for maximum undistorted output rather than maximum power output. In both procedures, the drain current

Receiving converter module uses point-to-point wiring and is mounted in small Pomona cast-aluminum chassis.



145-MHz output, and the second-harmonic was 45 dB down. I didn't have the opportunity at this time to balance the mixer, so it's conceivable the local oscillator could have been adjusted so that it was greater than 35 dB down. An appropriate antenna bandpass filter should significantly improve these figures.

Although I wouldn't recommend this project to a beginning vhf enthusiast, I can say I've had few problems with the project. I've built two complete transverters of the 2-watt variety, one 5-watt linear, and two 10-watt linears with the same good results. Both transverters are on the air and work very nicely.

In conclusion, I'd like to thank WA6COB and WA6VAB for their help on this project. Special thanks go to Ed Oxner of Siliconix, and Will Alexander, WA6RDZ, for their enthusiasm and technical support and to WA6RNC for the photography.

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2. George D. Hanchett, "Insulated-Gate Field-Effect Transistors in Oscillator Circuits," RCA Electronic Components and Devices, Somerville, New Jersey.
3. Harold T. McAleer, "Mixer Circuit Has Clean Output," General Radio Company, Engineering Department (reprinted from *Electronic Industries* October, 1960).

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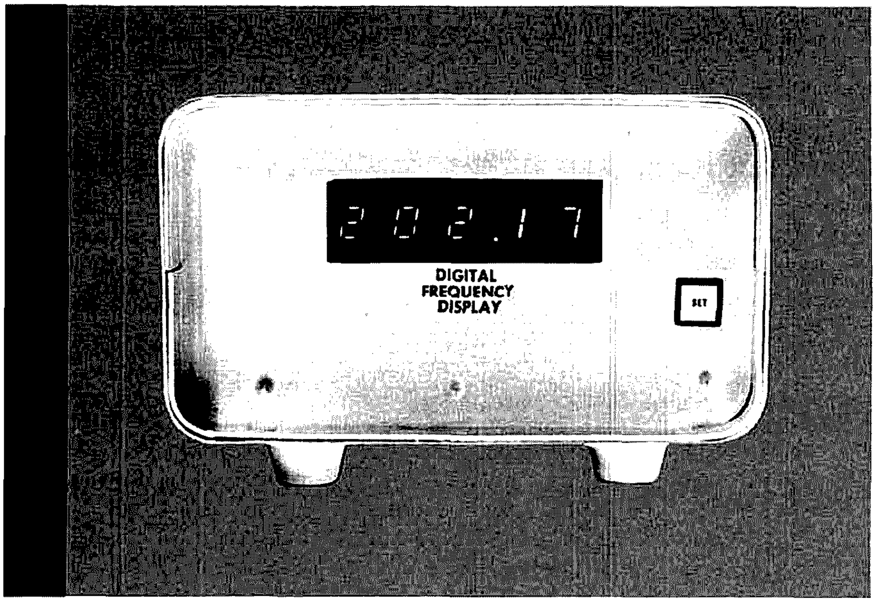


digital frequency display

for amateur communications equipment

Although designed for Heath SB-series equipment this compact frequency display is easily adapted to most amateur gear

During the past few years a number of frequency displays have been offered by manufacturers or described in the amateur literature.[1-3] I found that most had some undesirable feature: cost, complexity, or difficulty in setup and adjustment. The design I finally arrived at satisfied my needs and hopefully it will be useful to others.

Although this display was designed for the Heath SB series, it will work well with nearly any receiver, transmitter, or transceiver. Construction cost should be less than $50 even if you have a barren TTL junkbox. No exotic parts are required, so it should be no problem obtaining everything required from the popular parts distributors.

theory of operation

Let's begin by looking at a typical amateur transceiver, such as the Heathkit SB101, to see what's required to make a digital frequency display. **Fig. 1** shows a simplified block diagram of the SB101 in receive mode. As you can see, there are four signals to consider in computing the received signal frequency: the first heterodyne oscillator, the vfo, the carrier/product oscillator, and the audio output frequency. There are at least

By Bruce McNair, WB2NYK, 12 Marion Avenue, Howell, New Jersey 07731

three different ways of combining these signals to yield the correct frequency, and I've seen all of them used in the past.

The first method is to build a mixer to combine all the signals and come up with an rf signal to count. However, unless you go to great pains to shield everything, it's likely your frequency readout system will sneak enough signal back into the receiver to cause havoc.

proper choice of the digital counter. The 74192 programmable up/down BCD counter is a handy device to use in this and many other applications. This device has been described in other articles so I'll briefly outline its operation.

Frequency subtraction. As shown in fig. 2, this counter has *up/down* inputs and *carry/borrow* outputs. When a pulse appears on the *up* input, the counter advances to the next higher state (i.e., from state 7 to 8). Likewise, for the *down* input, the counter goes to the next-lower state. When the counter is in state 9 and an *up* signal comes along, the counter goes to state zero and signals the next stage, through the carry line, to advance. Corresponding events occur when the counter is in state zero and receives a *down* signal.

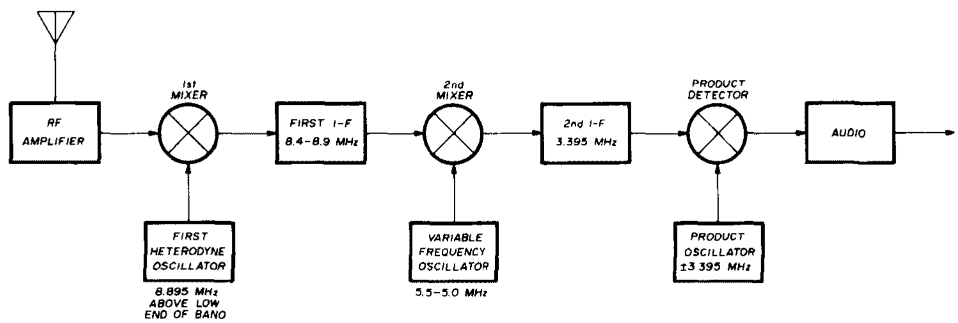


fig. 1. Simplified block diagram of the Heath SB101 in the receive mode showing signal sources that can be used for a frequency display. The vfo frequency is used as the variable to implement the frequency counter in the design described here.

The second approach is to combine the four signals digitally by using an up/down counter. The counter adds by counting upward and subtracts by counting downward. For any particular mixing scheme in a rig, this method is workable, but the counter must be reprogrammed for a different rig using a different combination of oscillator signals.

Finally, the third approach takes advantage of the fact that two of the four signals are crystal-oscillator signals and are fairly stable. These are the first heterodyne and carrier/product oscillators. The third signal is the audio output and, in most cases, can also be considered to be constant. This leaves only one variable – the vfo.

Realizing this, all you need do now is determine the vfo frequency at some given frequency and later determine the offset from the original vfo frequency. For example, if you measure the vfo at 4995 kHz when the receiver is tuned to 14.000 MHz, and then measure 5003 kHz when tuning an unknown frequency, you know the unknown frequency is 14.000 + (5.003 - 4.995) or 14.008 MHz. Remember that in the Heath SB series the vfo is tuned backward up the band (because the second i-f is the difference between the vfo and the first i-f).

implementing the theory

The next problem to solve is, how do you accomplish the subtraction of frequencies? Also, how do you determine and store the vfo frequency at the band edge? It turns out that both problems can be easily solved by the

Two other useful inputs are *clear* and *load.* Indepen-

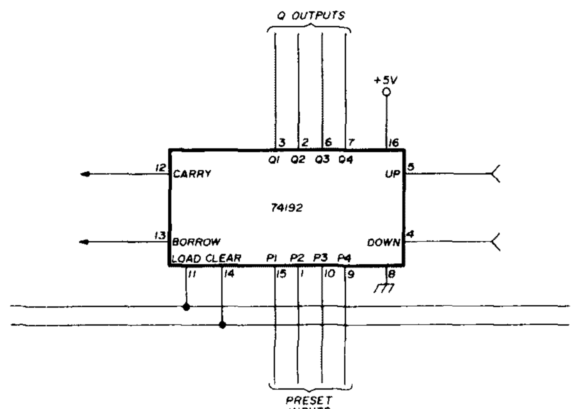


fig. 2. Heart of the frequency display is the 74192 programmable up/down BCD counter. UP/DOWN and CARRY/BORROW inputs and outputs are used to subtract frequencies and store the vfo frequency at the band edge in the counter display.

dently of whatever else the counter may be doing, when a pulse appears on the *clear* input, the 74192 goes to state zero. Also independent of other inputs, when a load signal is received the data present on the *preset* inputs are loaded as the new state of the counter. For example, if the 74192 were in state 7 and a load pulse

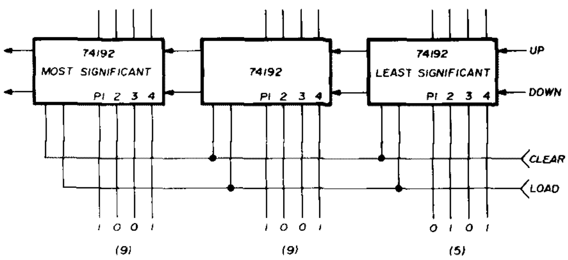


fig. 3. Three 74192s in cascade form a counter that will count to 999.

occurred, and if the *preset* inputs were 0101 (5 in decimal), the counter would go immediately to state 5.

There are certain restrictions in the use of this counter, but they are logical. Telling the counter to simultaneously count up and down is ambiguous; likewise, telling the counter to load a *preset* and *clear* at the same time does not make sense.

Now, how is this device useful? Suppose you had three 74192s cascaded to form a counter that will count to 999, as shown in fig. 3. Assume you initially loaded the counter with 995. Now you gate the vfo into the counter through the *down* input for 1 ms. Initially you are at 14.0 MHz so the vfo is at 4995 kHz. In 1 ms, 4995 pulses get through to the counter. Well, 995 (the initial state) minus 4995 (the number of *down* pulses) equals -4000. What state is the counter in at the end of the counting interval? Is it -4000? Not really, because the counter only remembers the three least-significant digits. The -4 showed up as four borrow pulses from the last stage. Therefore, you are tuned to 14.000 MHz and the counter displays 000. When you tune to 14.008 MHz, the vfo will be oscillating at 5003 kHz, and a 1-ms gating signal will admit 5003 pulses. The final state of the counter is now 995 - 5003 = -4008. Again, only the three least-significant digits are seen and you see 008, corresponding to 14.008 MHz. It's easy to see that a four-digit counter and a 10-ms gate would display to the nearest 0.1 kHz.

Calibrate sequence. It looks like the first question is now answered: to subtract frequencies you just load the counter with the *preset* for the band edge and count down. But now for the second problem. How do you determine the frequency at the band edge and save it for future presets? This question can be answered in terms of what you already have — a device that will start at zero or any preset number and count up or down from there. Normally you start at some *preset* and count down. What if you started at zero and counted up to some final state? If you then saved that final state for your new *preset*, everything would be solved. You would start at zero, count up to a final state, then count back to zero. This "calibrate" sequence could be initiated by pushing a button to activate the proper control logic. The vfo would be initially set to the low edge of the band (while tuning in the rig's 100-kHz calibrator), and the counter would display all zeros. As the vfo is tuned away from the band edge, the counter will read the operating frequency accurately.

Let's look at a diagram of one decade of the counter circuit required. **Fig. 4** illustrates what is needed. For those who can understand a timing diagram better than my word description, see **fig. 5**, which also shows the control signals the counter requires.

In fig. **6**, I've put the whole *counter* together in a block diagram. The *set* pushbutton initiates the calibrate sequence described above. Finally, fig. **7** shows what is required for control and timing logic. As described under

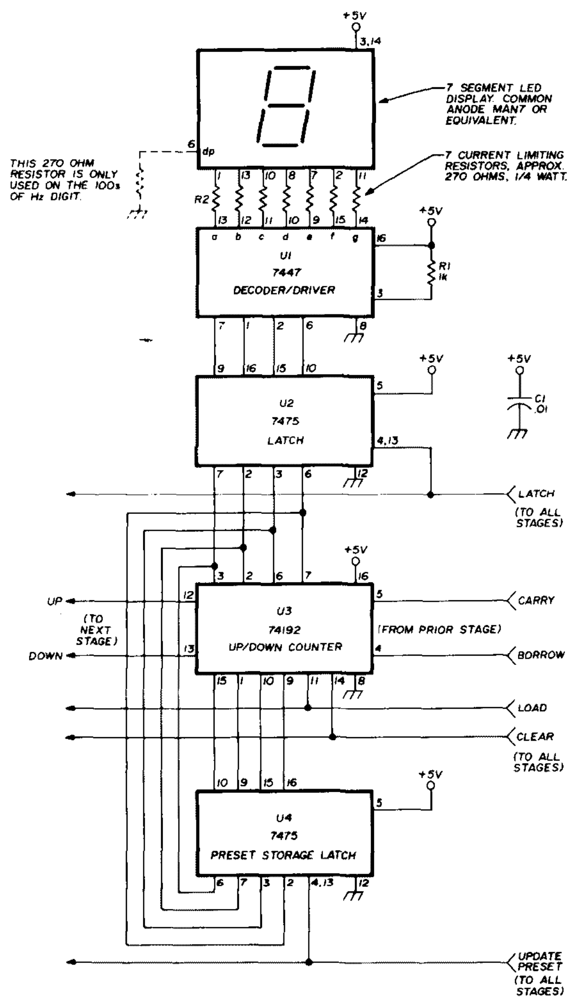


fig. 4. One decade of the counter used in the frequency display.

construction and adjustment, this board contains the only variable element: the crystal oscillator. As shown in fig. 7, two options are available for this oscillator. Initially I used a simple and reliable CMOS oscillator circuit. The circuit is, however, difficult to trim (I had to open up the crystal can to do so!). If you decide to use that doesn't tune backward, simply reverse the *up/down* lines that come out of the control board to the first decade.

construction

I chose to put all control logic including the time-base

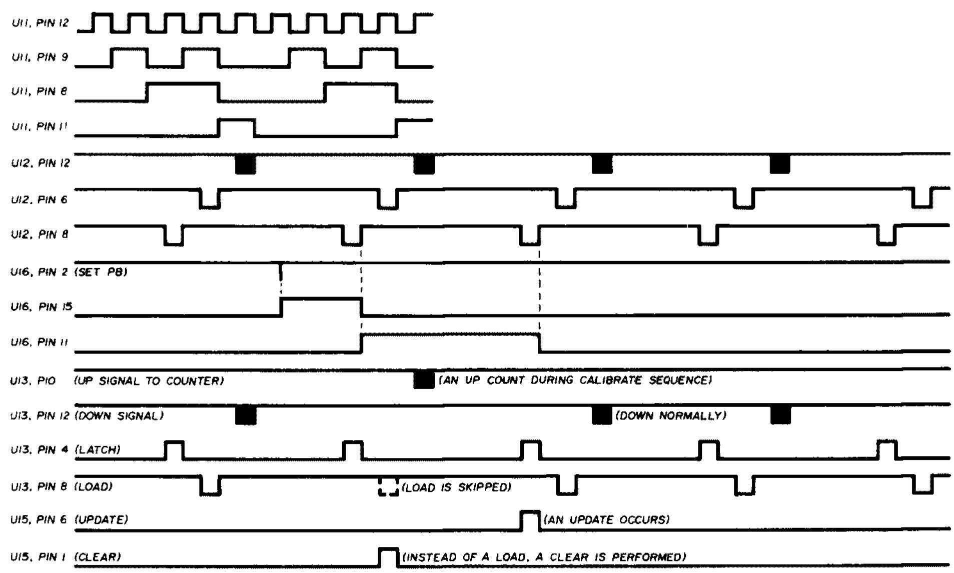


fig. 5. Timing diagram of the digital-frequency display including control signals.

the alternative oscillator shown on page 21, it will have to be external to the control board but will be much easier to adjust.

As described, the frequency display will operate with nearly any rig in which the vfo tunes backward and will indicate frequency within 10 Hz. If your rig uses a vfo oscillator and count-down circuitry on one board. Each decade of the counter occupies its own board. This method seemed to give the most compact layout when using printed-circuit boards. It also made it five times easier to design the counter board(s)! Fig. 8 gives the foil pattern used for the control board, and fig. 9 shows

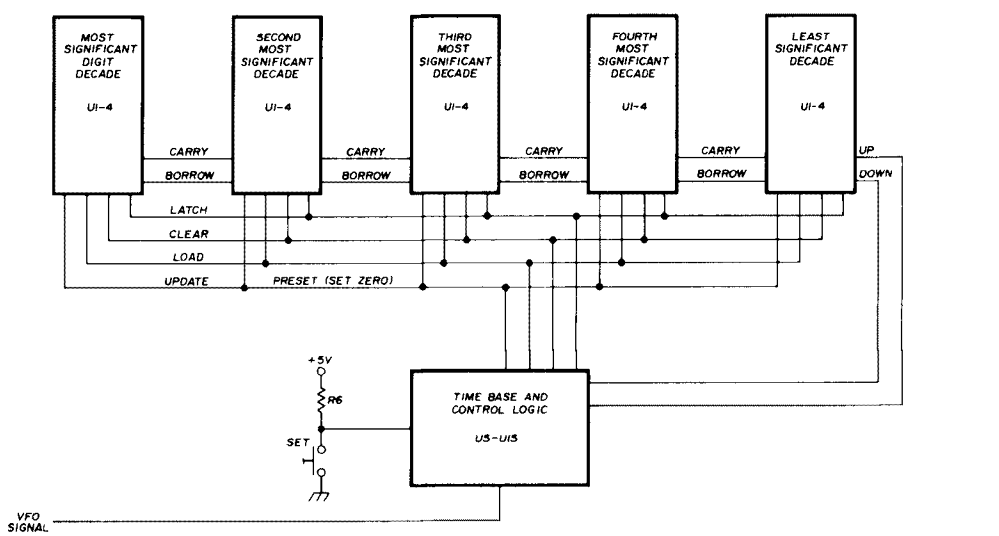


fig. 6. Block diagram of the frequency-display counter showing counter decades and time base and control-logic relationships.

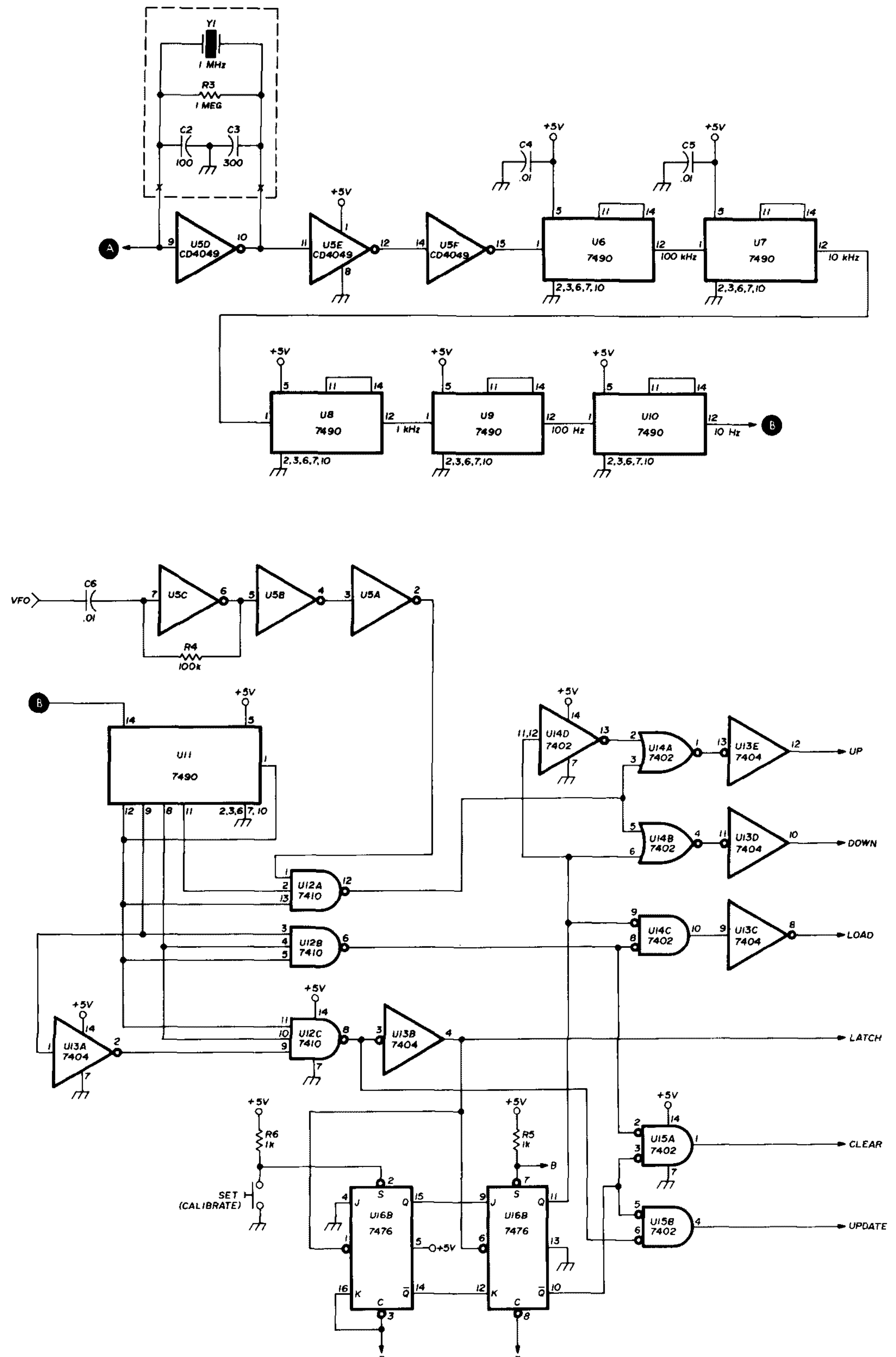
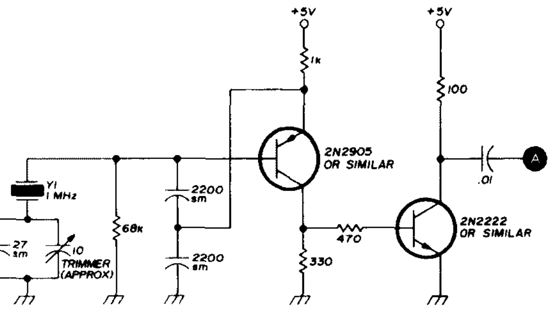


fig. 7. Control and timing logic. An alternative time-base oscillator, shown on the opposite page, must be external to the control board but is much easier to adjust than the circuit shown for U5.



ALTERNATIVE TIME-BASE OSCILLATOR

the part and jumper layout. To avoid the use of jumpers on the five decade boards, I used a double-sided board, illustrated in **fig. 10**, with component layout in **fig. 11**. Note that if you wish to avoid double-sided boards, the component side of this board is really just a number of jumpers and could be eliminated. I have tried to show my chassis layout in the photographs, but realize that almost *nothing* is critical about the layout.

George Oliva, WA2UOA, used an approach that may interest some people trying to duplicate this circuit. He used two separate chassis, one for all logic — hidden away under the rig and a tiny LMB chassis for the displays and *set* pushbutton, both sitting on top of the rig.

power supply

Since all circuitry is TTL (or CMOS running at TTL levels) the logic needs a fairly well regulated 5-volt supply. Current consumption, if you use standard 7400-series TTL, will be about 1.5 amperes, with each board (control and decade) drawing 220 to 230 mA. If you have or feel like buying some 74LS low-power Schottky logic (especially for the 74192s) the load on the supply will be eased considerably.

I decided to save my 74LS192s for a project requiring more speed and now live with a rather warm three-terminal regulator.[4] I chose to power all five decade boards from one 7805 regulator in a TO3 case mounted on the rear panel and to power the control board with its own 78M05 mounted on the board in a TO5 case. In both cases, unregulated dc was 11 volts dropped to 7 volts through a series resistor (R7 and R8 in **fig. 12**) to prevent the regulators from overheating because of a 6-volt internal drop. Make sure any three-terminal regulator you use is well decoupled on both the input and output. At least 0.1μF is required to prevent oscillation under load or at higher temperatures.

If you'd like to stay with a single regulator, it will be necessary to use an external series pass transistor to handle the current. All these ideas are shown in **fig. 12**.

installation

I hope I haven't given the impression that this digital-frequency display works only on receive — it was just easier to describe that way. For the Heath SB101 and most other sideband transceivers, the first heterodyne oscillator and the carrier/product oscillator are common to receive and transmit. With my SB101, I can read transmit and receive frequencies with an inboard or outboard vfo without any switching. This is the installation I describe, but any other should become obvious.

On receive, the vfo injection circuit looks like that in **fig. 13**. The point labelled A on one side of R221 is the best place to pick off the receive vfo signal, whether an internal or external vfo is used. On transmit, the circuit is as shown in **fig. 14**. The point labelled B on one side

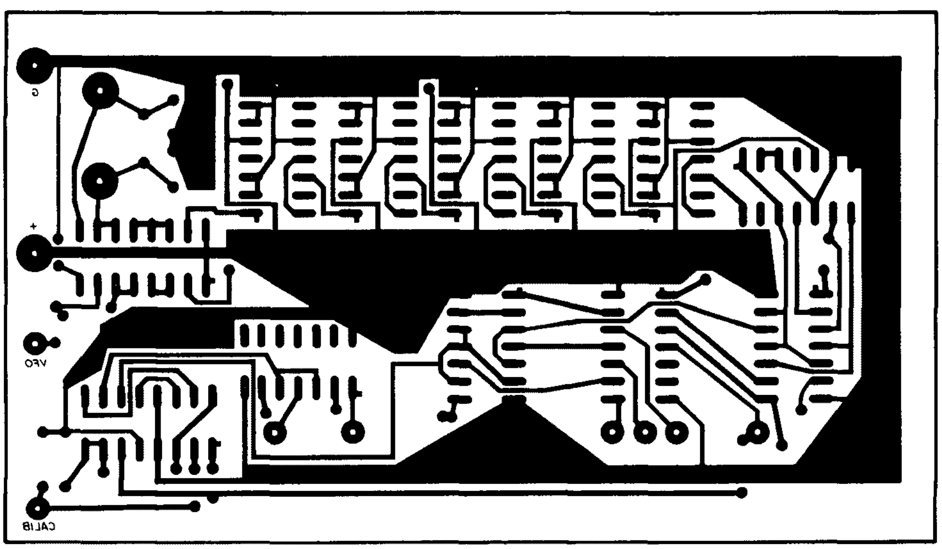


fig. 8. Foil side of the control board.

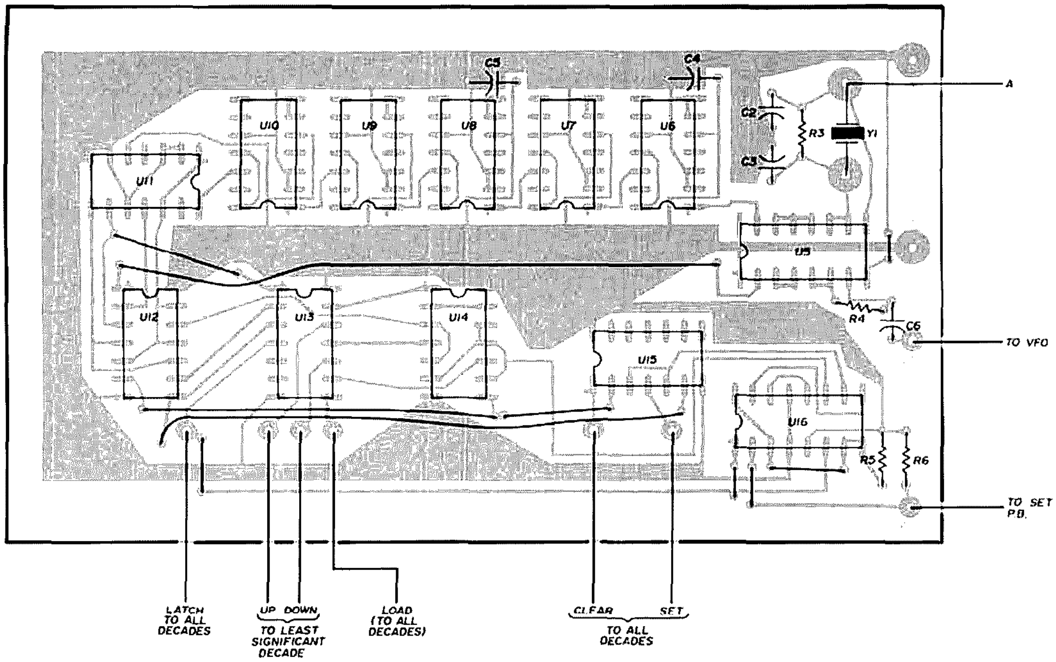


fig. 9. Component location and jumper layout for control board.

of R927 is the best place to take off the transmit vfo signal, also independent of the control mode.

To combine these two signals, I used the simple circuit of fig. 15. If you've built your own external vfo, or are using a different rig, make sure that power is applied only to one vfo at a time when in the transmit mode, or you may find yourself operating on two frequencies at the same time!

The CMOS circuit used for buffering the vfo signal should be adequate for most rigs. Mine has worked

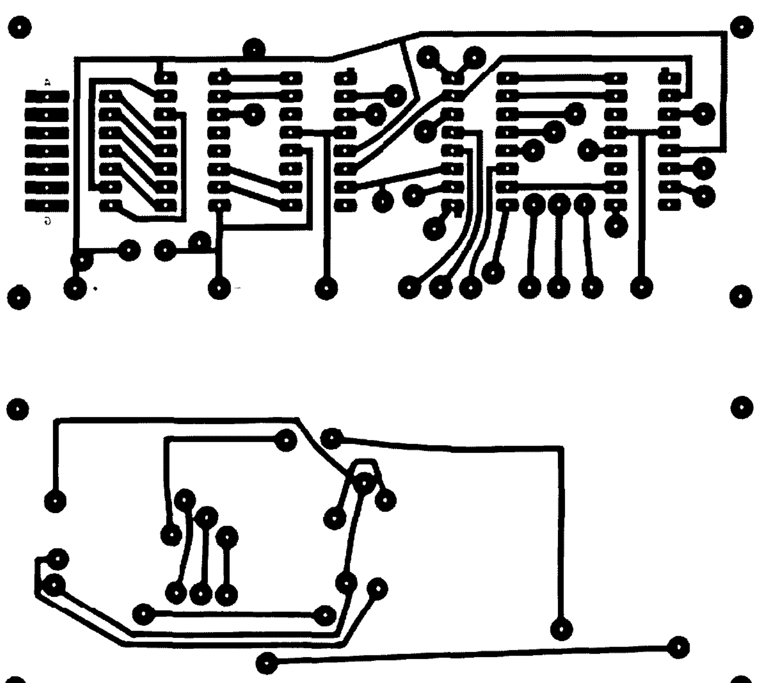
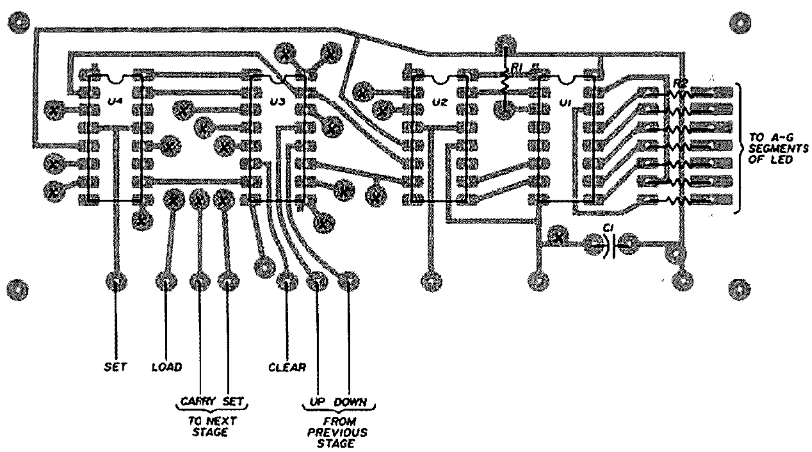


fig. 10. Double-sided board for the decade counter. Foil side is above, component side is below.

fig. 11. Part and solder plug locations for the decade counter board. Symbol X denotes location for connection between foil and component sides of the board. Traces on component side are not shown (see fig. 10); these circuit traces could be eliminated if desired (see text).

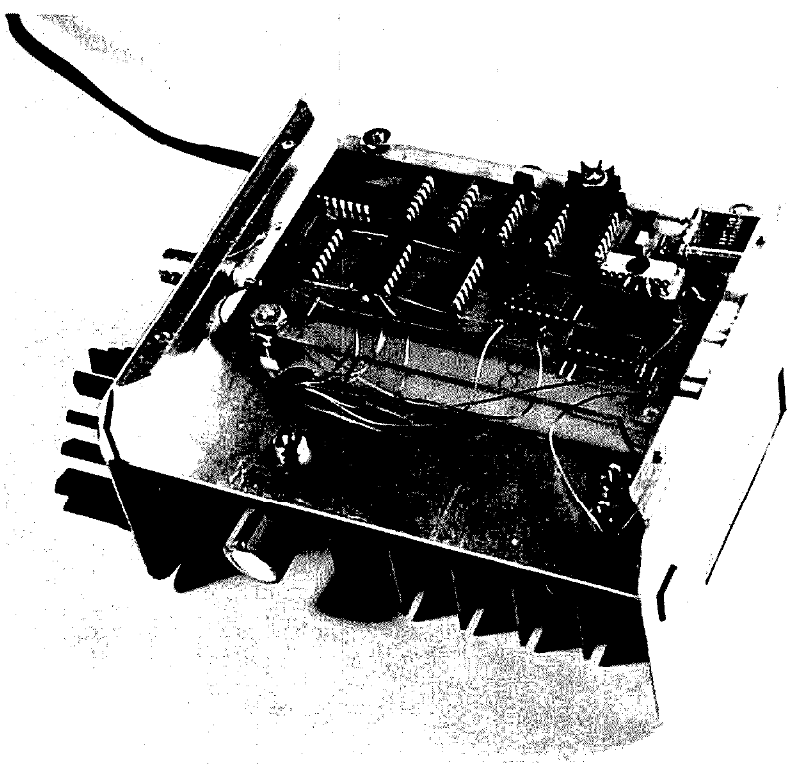


reliably with signal levels of 500 mV at 5 MHz and was good to 50 mV with the breadboard.

The only adjustment is in the crystal oscillator used for the frequency standard. To guarantee accuracy over the entire 500-kHz tuning range of the Heath SB series, when calibrated at one end, the 1-MHz oscillator must be within 5 Hz, which should be easy to ensure. For Collins equipment with a 200-kHz tuning range, the oscillator need be within 12 Hz.

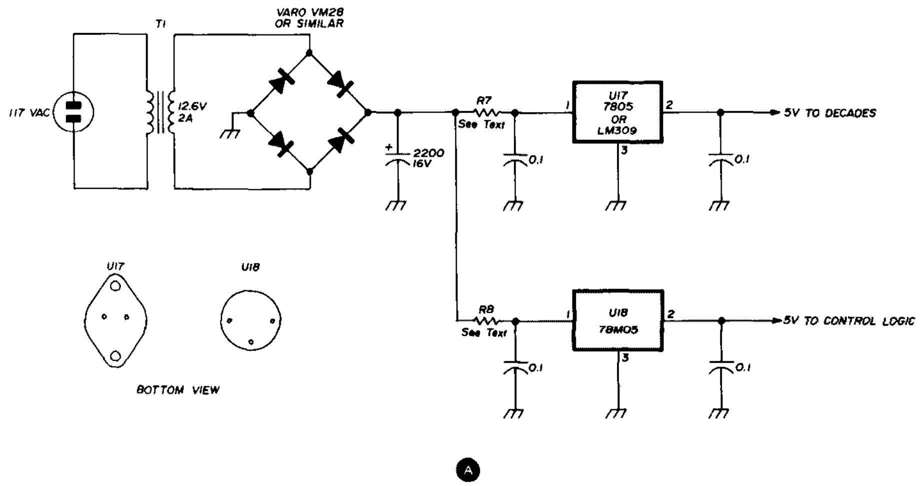
As mentioned above, if you use the onboard crystal oscillator, it will be necessary to open up the crystal (preferably in a HC6/U holder) and make pencil marks on the center of the crystal to lower its frequency. Do this while listening to WWV for a zero beat or while measuring with a frequency counter. An alternative method would be to zero beat an a-m broadcast station, although the frequency tolerance of these stations is rather poor for this purpose. To zero beat an a-m station, use the 10-kHz output of the 7490 divider. Whatever method of frequency measuring you use, you'll probably find it much easier if you have built the external 1-MHz oscillator that can be trimmed with a capacitor.

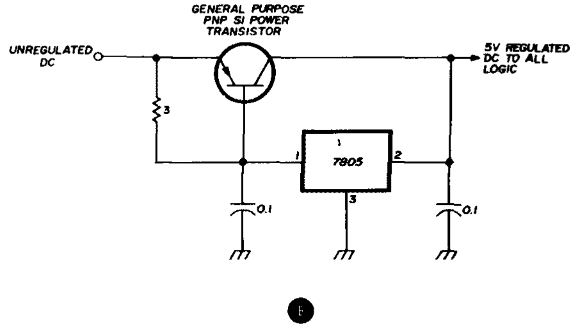
The use of this counter should be evident from the



Bottom view of the digital frequency display showing control logic and time-base circuits.

fig. 12. Power-supply suggestions for the frequency display. A separate 5-volt supply is shown in A; B shows a single supply using a series pass transistor to handle the current.





description of its theory of operation, but there are a few things I'd like to point out. First, when tuning CW, remember that the transmitter operates with 1-kHz offset in the carrier oscillator. Therefore, if you want to display both transmit and receive frequencies accurately, it will be necessary to tune in the calibrator with the same offset. Hopefully, this offset will be the point where the CW filter peaks up. For single-sideband operation, it will be necessary to zero beat the calibrator. Remember that the vfo and carrier generator are different for upper and lower sideband; you'll have to recalibrate when switching between them. This shouldn't cause too much inconvenience, since USB and LSB usually are not mixed within any given band. Finally, remember that it will be necessary to recalibrate when switching bands, since you are changing the first heterodyne oscillator crystals. On the same band, when operating in the same mode, recalibration normally will not be required.

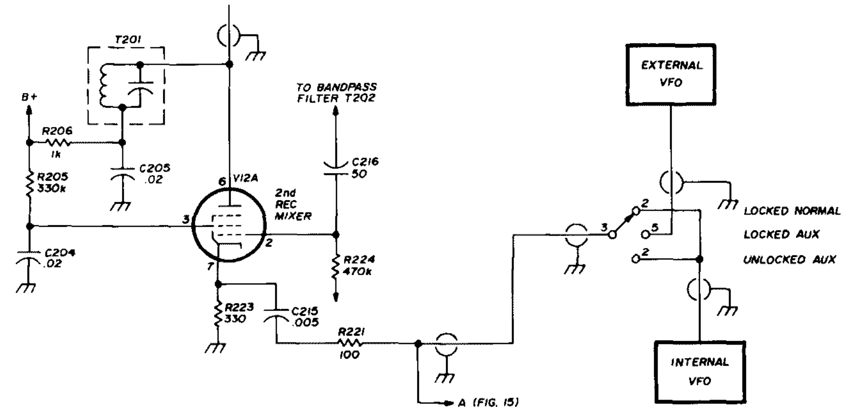


fig. 13. The Heath SB101 second receive mixer circuit showing where to obtain the receiver vfo signal.

possible modifications

Let me point out a few simple changes you might consider to make your digital frequency display more convenient and more versatile. First is changing from 10-Hz to 100-Hz readout. You may not feel that 10-Hz accuracy is needed in your station. This is actually the easiest change of all: just leave out the least-significant digit decade module and bypass one of the 7490s in the time base divider. The display will then be updated 10 times faster (10 times per second) and some savings will be realized (about $7 in construction cost and about 200 mA of current consumption). By the way, this change will also relax the requirements for time base accuracy; 1 MHz ±50 Hz will be sufficient.

The next possibility is the installation of a double-pole, double-throw switch in the *up/down* signal lines from the control board to the first decade. If this switch is rigged to reverse these lines, you can operate the display with forward and backward tuning vfos.

suggested parts sources

A large number of distributors sell the parts used in this project, many of whom advertise in *ham radio.* Two I've had experience with are James Electronics, Belmont, California and Solid State Systems, Columbia, Missouri.

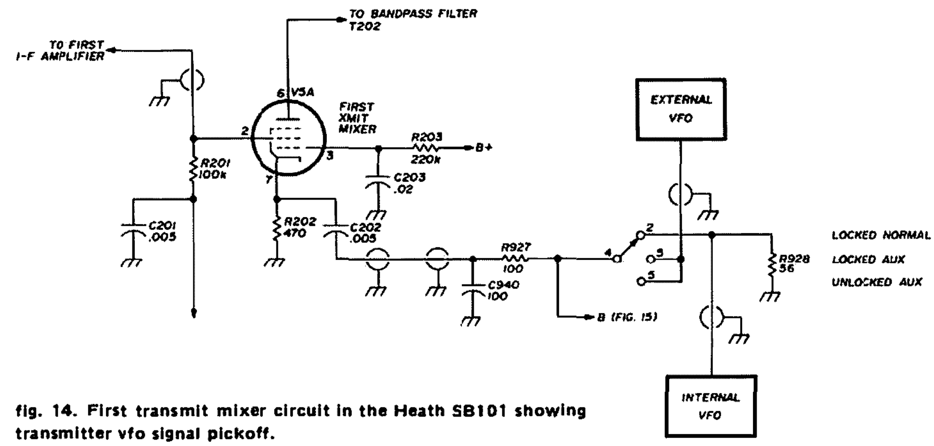
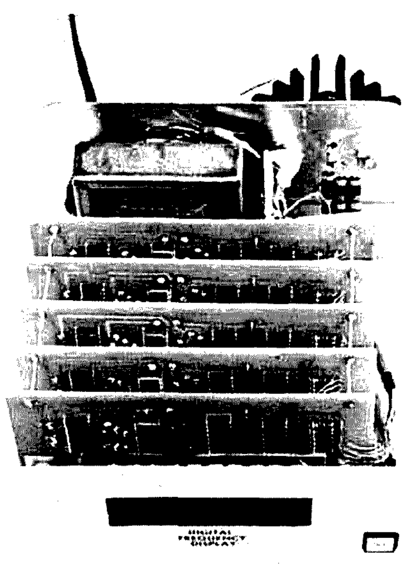


fig. 14. First transmit mixer circuit in the Heath SB101 showing transmitter vfo signal pickoff.

Finally, if you have installed the forward/backward switch, you now have a new piece of test equipment for the shack: a general-purpose frequency counter with direct readout to 1 MHz and 10-Hz accuracy. To make the counter even more versatile, you can build a preamp and bypass the CMOS buffer, yielding a counter usable to at least 20 MHz with the capability of measuring frequency drift and offset automatically (good for checking the drift characteristics of a new vfo).



Digital frequency display showing arrangement of decade counters.

All parts used in this project are stocked by at least one of these suppliers on a regular basis.

conclusion

I've enjoyed designing and building this frequency readout system and hope I've given enough information so that others can duplicate it. I'd be interested in

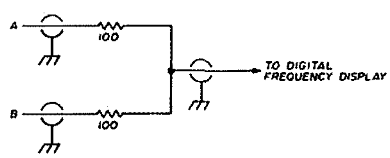


fig. 15. Circuit for combining the signals obtained in figs. 13 and 14.

hearing of any difficulties encountered. I'd also like to hear of any improvements or changes others may come up with.

acknowledgements

I'd like to thank George Oliva, WA2UOA, for his assistance with the photographs and Glenn Williman, WB2DHG, for his constant and occasionally useful criticism.

references

1. Jon Hagen, W7URZ, "A Simple Frequency Counter for Receivers," *QST*, December, 1972, page 11.
2. Kenneth MacLeish, W1EO, "Frequency Counter for the Amateur Station," *QST*, October, 1970, page 15.
3. *Recent Equipment,* "Heath Frequency Display Model SB650," *QST*, August, 1972, page 56.

ham radio

the Accu-Mill

a keyboard interface for the Accu-Keyer

With this circuit and an ASCII-encoded keyboard connected to an Accu-Keyer you can send perfect Morse at 25 words per minute or better

More than six thousand Accu-Keyers have been built worldwide since the circuit was introduced in the August, 1973, issue of *QST*.[1] The Accu-Mill connects to the Accu-Keyer or to an Accu-Keyer with memory using simple circuitry. With it you can key in your message from a typewriter keyboard, then start sending with your paddle without throwing switches – great for contest operating. The basic Accu-Mill circuit requires a computer-type ASCII-encoded keyboard (many are beginning to show up in the surplus market). You'll also need an Accu-Keyer with an extra-large power supply and a negative 12-volt supply. The following features have been designed into the Accu-Mill:

1. Sixty-four or 128 characters of buffering.
2. Usable with the Accu-Keyer or Accu-Keyer with memory, including provision for external paddles.
3. "Buffer Full" light.
4. Nonbuffered operation with repeating keys and nonbuffered operation with non-repeating keys, both switch selected.
5. Speed control by the Accu-Keyer.
6. Easy interface with keyboards having standard or inverted outputs.
7. Only one circuit board.

The Accu-Mill was also designed so that other functions can be added such as:

1. RTTY
2. Radio ASCII when approved.
3. Counter with digital display to let you know how much buffer has been used.
4. A back space function to allow correction of keystroke errors.
5. Three programmable "vectored" message memories that work with the keyboard buffers to allow automatic insertions in a programmed contest report, so that you can type variables such as callsign, RST and number, for example, and the machine will make insertions while you update the log.

All these extras are now on paper and should be available soon. The memory and backspace features are particularly exciting, because they allow a 10-wpm typist who makes lots of mistakes (like me) to operate in a contest at 25-wpm or better.

logic description

The basic Morse board is presented here. The board uses easy-to-obtain 7400-series TTL devices except for the buffers, which are Fairchild OptiMOS devices; and the read-only memories (ROMs), which usually have the 8200-series TTL numbering (8223, 82S23, and 82S123 tri-state), but also have the 7400 designation of 74188.

Most of the devices have standard totem-pole outputs, which provide logic levels of 0 and 1. Some have open-collector outputs, which provide logic levels of 0 and not-0. These special types are used to connect to the outside world in places where either another output or another device (such as paddles) will operate an input. For proper operation we provide the logic 1 level with an external pull-up resistor. This resistor provides 5 volts at low current when the device is in the not-0 state. When the device is in the 0 state, the open-collector output shunts the voltage from the resistor, and a 0 logic level occurs.

The buffers are 3341 types: first-in first-out (FIFO) shift registers, the same as used in every popular buffered Morse or RTTY keyboard. Here's how they work: A short *shift-in* pulse commands data to enter. Data goes into the input and "falls through" to the last unoccupied slot of 64 positions. Data is output by a *shift-out* pulse of any length. By analogy, imagine a long, tilted gutter with a man at the top dumping in tennis balls as fast as he can and another man at the bottom, who takes them out when he needs them. As long as the man at the top is faster than the man at the bottom, there will always be a supply of tennis balls in the gutter. That's how

By William E. Smith III, WN9OVY, 1880 Bonnie Lane, Apartment 201, Hoffman-Estates. Illinois 60194

Fairchild explains it. If you can type fast enough you'll get ahead of the output and your sending will be very smooth, or you can load up the registers and go out for coffee while the machine works your contact.

The read-only memories are of the field programmable type. You can easily program them to your code with a very simple circuit. They are used here as look-up tables. Each time a binary address is placed on their inputs, the ROMs "look up" the proper code associated with that binary address and present that code to their output lines. The 74151s are data selectors or multiplexers; they act something like single-pole, 8-throw switches, and scan the ROM outputs.

functional description

Buffer. The Accu-Mill logic diagram is shown in fig. 1. U1 is used only for ASCII keyboards with inverted outputs. U2 and U3 condition the signal from the keyboard that says a key has been pressed (KP). U2 and U3 provide the very short pulse needed for the buffer *shift-in* command and the control needed to disable the buffer for non-buffered operation.

U10-U13 are the 3341 buffers. Each buffer contains a channel four bits wide. We need a six-bit data channel, so the buffer is configured as two channels of data buffering, each four bits wide (two unused bits). It is probable that each channel will operate at a different natural speed, which would result in bad data output, so U14 synchronizes the two channels.

The buffers retain stable data on the output pins after *shift-out* returns low. If stable data were to remain on the input to the next stage, the keyer would lock up on one character. U15, U16 are NAND gates. When *shift-out* returns low, the gates stop data, and the next stage gets an acceptable signal (binary 111111), which prevents lockup.

So far we've looked at the first stage of the Accu-Mill. In this stage we input data from the keyboard and subject it to buffering so that we can get ahead of the transmitted output and type at our own speed. The next stage contains the magic that makes Morse code.

Morse-converter. This stage includes seven devices. U17-U20 are the read-only memories, which contain the ASCII-to-Morse conversion logic. U21, U22 are data selectors, which read the Morse code. U23 is the control counter, which drives the data selectors.

Data at the buffer output is presented to the memory address lines. Since each key has a different binary number, each key addresses a different place in memory, and the memories output a different Morse code bit for each key pressed. The Morse information appears at the memory outputs in parallel form: all dots and dashes appear at once. U21, U22 are data selectors, which put the Morse information in serial form. The data selectors start by looking at Morse bit one and sequentially scan all output bits until the character is complete. The memories are wired so that U17, U18 are dot memories, and U19, U20 are dash memories.

Accu-Keyer interface. This stage includes U24, U25. These gates provide control signals and interface to the Accu-Keyer. U24A, U24B replace the paddles. The changes in the Accu-Keyer called for in the wiring of the paddle jack (fig. 1) will greatly increase rf noise immunity in the Accu-Keyer. The use of 74132 Schmitt triggers in place of U1, U2 in the Accu-Keyer should cure even the most severe rf problems. Similarly, a 7414 hex Schmitt trigger could be used in high-noise environments in place of U1 in the Accu-Mill.

data flow

The letter A. The letter A is a good example to use in describing data flow through the Accu-Mill, because it has just one dot and one dash. So press A. The keyboard generates 1000001, a seven-bit binary number. But a six-bit number is sufficient to describe all 64 possible characters in memory, so ignore bit seven and leave that wire disconnected.

The six-bit code 000001 is presented to the buffer stage. KP generates *shift-in*. 000001 is entered into the buffer and falls through to the output. U14 senses valid output data and opens gates U15, U16, which invert the data. Now the data is 111110, a six-bit code that is presented to the memory section.

Each memory has only five address lines, so it would appear that there's another unused bit. The extra bit, bit six, is used to select which 32 of the 64 characters is being addressed, letters or numbers. Bit six, which is high (1), is fed to, but does not enable, memories U17, U19, which are programmed for numbers and punctuation. Bit six is also inverted by U25A, and fed to U18, U20. This low (0) signal enables U18, U20, and the memory is ready to make a letter.

The remaining five bits, 11110, address U18 (dot) and U20 (dash) to the places where Morse information necessary to make A is stored. Dot memory U18 outputs 10000000; dash memory U20 outputs 01000000. Data selectors U20, U22 are looking at position one (far left). Dot data selector U21 sees a high bit in position one and causes inverter U24A to ground the dot input of the Accu-Keyer, and a dot is sent. On the dot falling edge, control counter U23 is clocked up by a signal fed back from the Accu-Keyer output.

The output count of U23 changes and drives both data selectors up to the next output lines in the memory. Now dash data selector U22 sees a high bit in position two and causes inverter U24B to ground the dash input of the Accu-Keyer, and a dash is sent. Again control counter U23 is clocked up, and the data selectors are driven up to position three. Neither sees a high bit, so neither sends a dot or a dash. The Accu-Keyer assumes end-of-character and sends a character space.

Spacing considerations. A special condition exists for the seven-baud word space. The keyboard has a space bar, but the Accu-Keyer is capable of sending character-spaces only – not word spaces, so the Accu-Keyer needs to be tricked. A word space always follows a character which has a character space. The character space will be the first three of the seven-baud word space. The space bar will be programmed to send the letter E through the Accu-Keyer. The letter E is one baud, which totals seven baud with the previous character space.

To avoid letting the E get to the transmitter, the

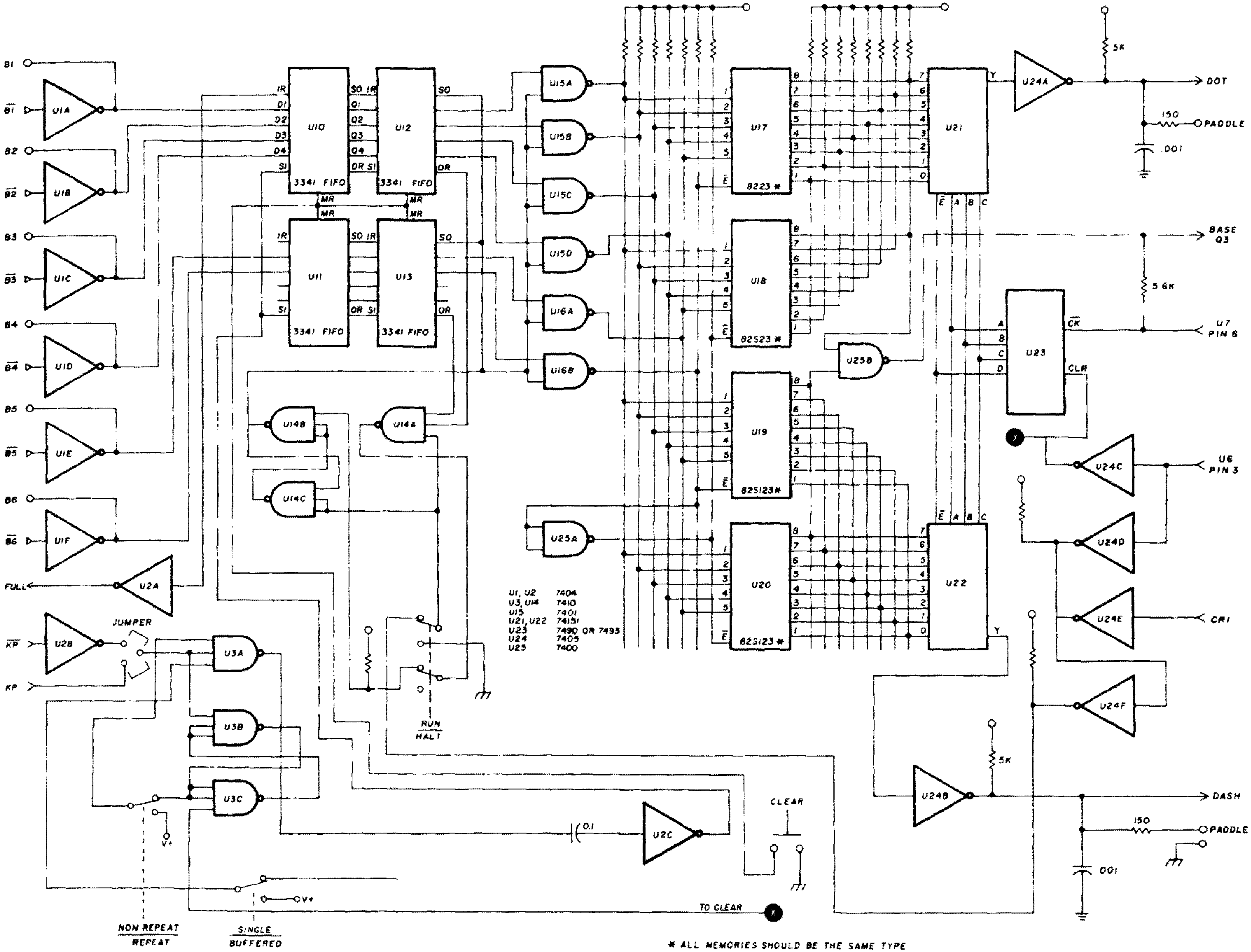


fig. 1. Logic diagram for the Accu-Mill. Shown are methods for connecting three different PROM ships, though all should be of the same type. A complete memory programming chart is supplied with the circuit hoard, or 82S123s are available preprogrammed. In this circuit all resistors are 2200 to 3300 ohms, ¼ watt, terminated with V+ unless indicated otherwise. All power-supply connections are assumed, as is proper bypassing of V+. Numbered wires are for logical reference, not pin numbers. Keyboard connections are on the left, Accu-Keyer connections are on the right.

memory is programmed differently for this special E to make it seem transparent at the transmitter. The dot memory will be 10000001; dash memory will be 00000001. Note that the first bit (left) in the dot memory will send a dot; the second bit is empty, which signals end of character. A character space is sent, the counter is cleared, and the next character begins. Note also that the data selectors never get to position eight to read the extra ones – this is the trick. These two extra bits are fed to NAND gate U25B. Its output goes low and swamps the keying signal before it gets to output transistor Q3 or Q5. As far as the Accu-Keyer is concerned the letter E was sent, but it never gets to the transmitter.

Using the same logic, a 36-baud pause can also be generated by programming a dot, 00000001, and a dash, 11111111. The keyer sends --- --- --- --- --- --- --- - ---. Including the character spaces, that's 36 baud of data, but it never gets to the transmitter. This feature can be put on the # key. The other upper case keys are used for signals such as $\overline{AR}$, $\overline{SK}$, $\overline{KN}$. The only key that cannot be used is the @ key. This space must be programmed in memory with all zeros since this space has the address 111111, which is the same as the resting code put out by U15, U16 between characters. If this space is programmed, the keyer will lock up. The @ will halt the keyboard when pressed. This may be cleared by the halt switch.

A ready made PC board is available for the basic circuit of the Accu-Mill.* It's double-sided and measures 3 x 8 inches (76x203mm), so it should fit in just about any enclosure. A parts-placement guide and step-by-step instructions are supplied with each board. Buffers and memories are also available. This project should not be attempted on vector or perf board.

Several companies sell ASCII-encoded keyboards. These keyboards are usually used or surplus items, but one company, at least, sells inexpensive new kits. New or used, as long as the keyboard works and looks good you're in fine shape. Many keyboards have square keys with no spaces between them, which makes the enclosure easy to build since you cut only one large hole instead of many small ones. Rf has never been a problem for me, but it makes good sense to keep the wiring clean.

Interface to the Accu-Keyer is easy and requires only about six wires. The Accu-Keyer controls the keyboard speed, so you may want to move the Accu-Keyer into the keyboard enclosure and mount your paddle jack on the side of the enclosure. If your Accu-Keyer has a keying monitor, remove the RC network from pin 4 of the 555 timer and connect pin 4 directly to Q3 or Q5 base. C4, C5, R11, and R12 on the Accu-Keyer inputs should also be removed and placed on the paddle jack, as indicated on the schematic.

U1 of the Accu-Mill is used only for ASCII keyboards with inverted outputs. To test for this condition just power-up the keyboard and press A. Test the outputs. Standard boards have bits 7 and 1 high (more than 2.4 volts); in this case don't use U1. Inverted boards have bits 7 and 1 low and everything else high; in this case use U1.

A note of caution: keyboards are available with codes such as BCDIC, EBCDIC, and Hollerith. These keyboards can be used but the programming in the memories is much different. A keyboard with separate contacts for each key is suitable only if an ASCII encoder is added to it.

The 3341 buffers are MOS devices and may be sensitive to static charges. Install these last and handle them carefully. Use two 3341s for 64 characters of buffering, or all four for 128 characters of buffering. It's possible to build either a nonbuffered board by using jumpers or a board with an outrageous amount of buffering (like 512 characters or more), but neither serves much useful purpose.

The memories can be the obsolete 8223 or the newer 82S23. Both require pull-up resistors (¼ watt, 2.2k). The very new 82S123 tri-state devices don't require resistors. Fig. 1 shows how to handle each type. A complete memory programming chart is supplied with the circuit board, or 82S123s are available preprogrammed.

IC sockets cannot be used since this is a double-sided board without plated-through holes. This type of board requires some soldering on the top side, and the plastic body of the IC socket will be in the way. Use Molex pins or solder the ICs directly.

In some places this circuit board has conductors running between adjacent pins, which means that the circuit pads are very small. A very fine-tipped soldering pencil is mandatory. Try grinding a spare tip down to a 1/32 x 1/2 inch (0.8x13mm) taper. Use very small-diameter solder. A soldering gun will ruin the board.

The 5-volt power supply should be capable of at least 2 amps. Use a 12.6-volt transformer, a bridge rectifier, and a zener diode with pass transistor and heatsink. Filter it well and use a transformer with about a 3 to 4 amp secondary rating.

The minus 12-volt supply should be capable of supplying 200 mA. Use a transformer with a 600-mA secondary rating, a bridge rectifier, and a zener with a current-limiting resistor. Filter the supply with 2500 μF at 25 Vdc ahead of the resistor and a 1 μF tantalum and 0.001 μF disc ceramic after the resistor.

The enclosure can be made of aluminum, wood, or even plexiglass. A flat piece of plexiglass with switches and keyboard mounted on top of a plexiglass or wooden box shaped like a typewriter looks good.

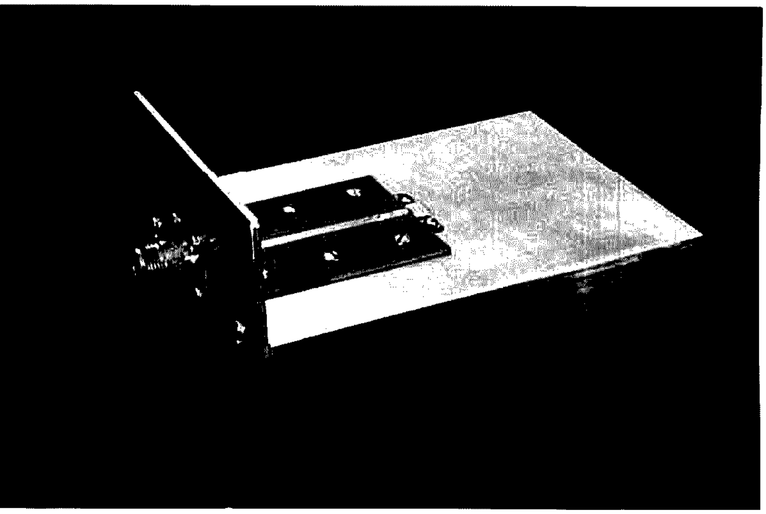
To operate the Accu-Mill, turn it on, select buffered or nonbuffered operation, and start typing. To use the paddle, just plug it in and wait for the Accu-Mill to finish, then start sending. You can change over without throwing any switches. The paddle may remain connected if desired. The HALT switch may be used at any time and will stop the machine at the end of the present character.

*An etched and drilled, double-sided, glass-epoxy circuit board is available for $15 from W.E. Smith, Post Office Box 544, Hoffman Estates, Illinois 60194. *Prices for 3341 buffers and 82S123 memories (preprogrammed) are available upon request.*

reference

1. James M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.

ham radio



150-watt uhf dummy load

Details of a 50-ohm, 150-watt dummy load that exhibits less than 1.2:1 vswr through 1500 MHz — total parts cost is less than $40

Although most amateurs have a dummy load that is suitable for use on the high-frequencies, most low-cost, high-power dummy loads are quite reactive at vhf and present anything but a 1:1 vswr. Occasionally good quality commercial loads are available on the surplus market at reasonable prices, but off-the-shelf vhf dummy loads that will handle 100 watts or more are priced out of the reach of most amateurs. The rf load described here, which is rated at 150 watts, and provides less than 1.2:1 vswr through 1500 MHz, can be put together for less than $40.

The heart of this unit is the CTC TA150-50 microstrip termination which looks much like a high-power uhf transistor. The device, which is only about 1/4-inch (6.5mm) wide and an inch (25.5mm) long, is attached to an aluminum heatsink as shown in the photographs. The input coaxial connector is connected to the load through a 50-ohm microstripline that allows close contact to the termination (fig. 1).

construction

Construction of the 150-watt load is simple, but a moderate amount of care is required to assure good uhf performance. The first step is to prepare the heatsink by milling a small indentation, about 0.1 inch (2.5mm) deep, in the center of the heatsink as shown in fig. 2. This allows the lead of the TA150-50 termination to be at the proper height to match the microstrip feedline. Installing the TA150-50 so its input lead is too high or too low may break the lead or crack the ceramic insulation. The correct mounting is shown in fig. 3.

Make sure that the indentation milled into the heatsink is reasonably smooth and flat. Any irregularities in the mounting surface for the TA150-50 termination, or bowing of the heatsink, may crack the BeO ceramic when the mounting screws are tightened. An imperfect mounting surface can also result in poor thermal transfer characteristics which will lower the power rating of the completed dummy load.

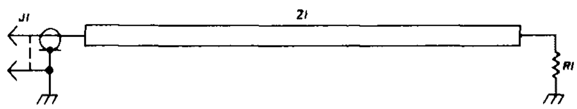
Although Teflon-glass circuit board was used in the author's model shown in the photographs, epoxy-glass circuit-board can be used if you do not expect to use the load above 300 MHz. This is not because of the higher loss of epoxy-glass circuit board (which is good to over 1000 MHz), but because of variations in thickness from one manufacturer to another which affect the impedance of the microstrip transmission line. In addition, voids in epoxy-glass circuit board can cause difficulties. If all you have is G-10 circuit board, by all means use it, but don't be surprised if the vswr above 300 MHz is substantially greater than 1:1.

Circuit-board dimensions for both materials are given in fig. 1. After the circuit board is etched and cut to size, six holes are punched in the board for mounting on the heatsink (see fig. 4). Do not use less than six mounting

By Lance Wilson, WB6QXF, Communications Transistor Corporation, 301 Industrial Way, San Carlos, California 94070

screws because they provide the ground for the 50-ohm microstrip feedline.

Temporarily place the termination in the indentation and mark the position of the two flange-mounting holes. Remove the termination and drill and tap the heatsink for 4-40 (M3) screws. Temporarily install the termina-



J1 UG-58/U type-N flange mount chassis connector

R1 CTC TA150-50 50-ohm, 150-watt microstrip termination (available from Webster Radio, 2602 East Ashlan, Department H, Fresno, California 93726)

Z1 50-ohm microstripline, 2.9" (7.4cm) long, width determined by board type as follows:
1/16" (1.5mm) Teflon-glass circuit board:
0.173" (4.4mm) wide
1/16" (1.5mm) fiberglass-epoxy board:
0.107" (2.7mm) wide

fig. 1. Schematic diagram of the 150-watt dummy load which exhibits less than 1.2:1 vswr from dc through 1500 MHz. The CTC TA150-50 rf termination must be mounted on a heatsink as shown in the photographs.

tion and line up the microstrip circuit board with the lead of the TA150-50. Position the board as close as possible to the termination and mark the six mounting holes for the circuit board. Remove the board and drill and tap the six locations for 6-32 (M3.5) screws.

Before permanently installing the termination and the microstripline, carefully clean the heatsink to make sure it is completely free of any oil or grease. Apply a small

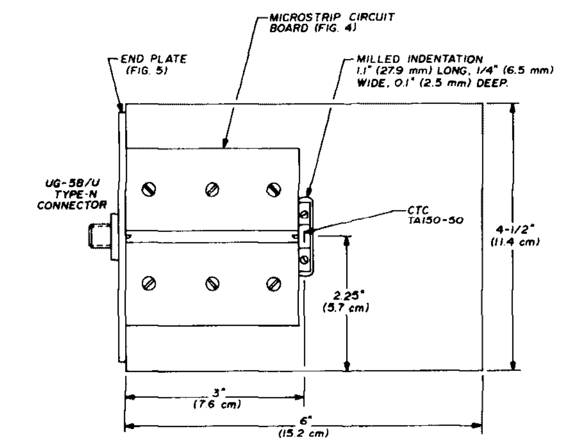


fig. 2. Heatsink layout for the 50-ohm 150-watt dummy load.

amount of zinc-based heatsink compound on the flange of the TA150-50, place it in the heatsink indentation, and install the two 4-40 (M3) screws. Alternately tighten each screw, but don't use too much torque or you may strip the threads in the heatsink. Now install the microstrip circuit board and the six 6-32 (M3.5) mounting screws.

The type-N UG-58/U chassis connector is attached to

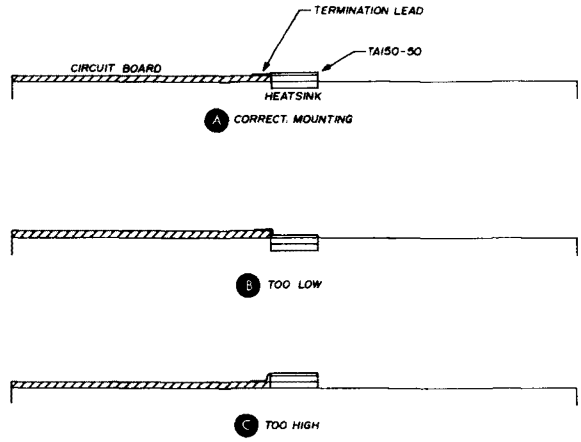


fig. 3. Correct installation of the TA150-50 is shown at (A). If the milled indentation in the heatsink is too deep (B) or too shallow (C), the termination lead will not match the height of the microstripline and may result in damage to the device. Correct depth is 0.1" (2.5mm).

an 1/8-inch (3mm) thick piece of aluminum stock, 4½ inches (11.4cm) wide and 2 inches (5.1cm) high (see fig. 5). Two holes in the end plate are used to mount it on the heatsink. The matching holes in the heatsink are drilled and tapped for 6-32 (M3.5) screws. Slightly elongating the two mounting holes in the end plate makes it easy to align the center pin of the coaxial connector with the microstripline. The connector is installed on the end plate with short 4-40 (M3) screws.

Install the end-plate assembly on the heatsink and adjust the position of the plate until the center pin of the coaxial connector rests on the top of the microstripline. Tighten the two 6-32 (M3.5) mounting screws and solder the connector to the microstripline. Now solder the TA150-50 lead to the microstripline. This completes the assembly.

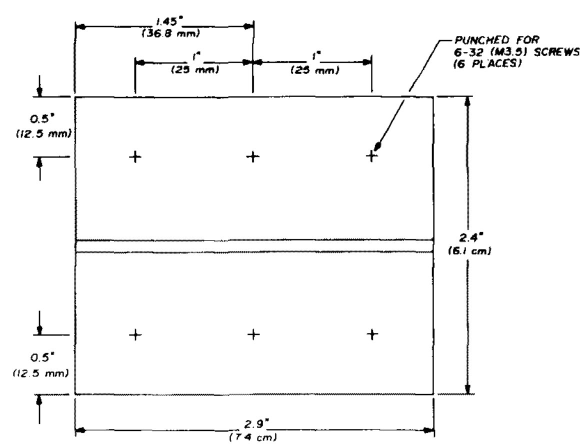
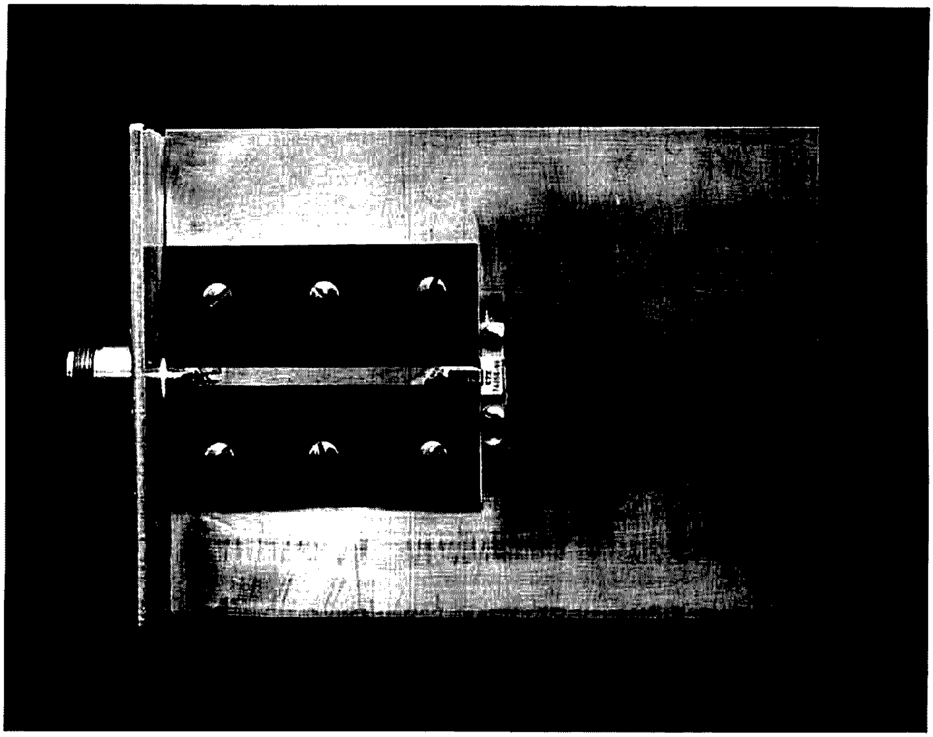


fig. 4. Drilling layout for the microstrip circuit board. Six mounting screws are required to provide good electrical contact with the heatsink. Printed-circuit layout is for 1/16" (1.5mm) double-clad, Teflon-epoxy circuit board, which is recommended for this application. Dimensions for fiberglass-epoxy board are listed in fig. 1.



Construction details of the 150-watt uhf dummy load.

checkout and performance

Connect an rf power source to an accurate vhf swr meter and the completed rf load as shown in fig. 6. Make sure the swr meter is accurate at the frequency of the power source (a Bird model 43 *Thruline* rf directional wattmeter with the correct plug-in element is a good choice). Gradually apply power to the rf load and look for any indication of high swr — this could indicate a poor solder joint, a short, or other assembly problem. If everything looks okay, go ahead and increase power up to the 150-watt maximum.

The maximum case temperature of the TA150-50 is 100°C (212°F). Since case temperature is difficult to measure, heatsink temperature can be used as an indication of safe operation. If a good grade of thermal compound is used when mounting the TA150-50, this will provide a case-to-heatsink thermal resistance of 0.3°C per watt.[1] This means that, at 150 watts dissipation, the heatsink temperature will be 55°C (131°F). If you cannot measure the heatsink temperature or are in doubt, force-air cool it (that's what I did). At lower power levels, 35 watts or less, the heatsink should be okay for *continuous* use without any external cooling.

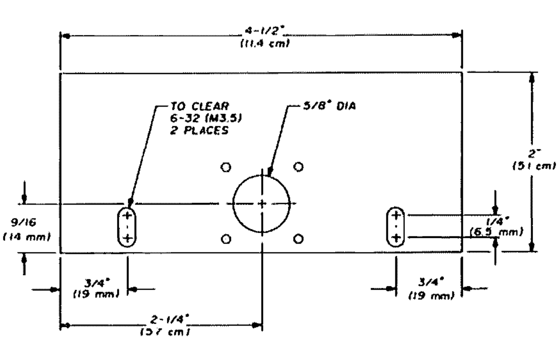


fig. 5. Layout of the end plate. The UG-58/U chassis connector is mounted with short 4-40 (M3) screws. Elongated mounting holes make it easy to align the center pin of the connector with the microstripline.

fig. 6. Test setup for checking the performance of the 150-watt vhf dummy load. Swr meter should be accurate at the frequency of interest (Bird model 43 with appropriate slug is recommended).

Note that the power limitation of the TA150-50 is 150 watts *maximum* — exceeding this rating, for either CW *or* PEP, may damage the termination.

The measured vswr of this dummy load is less than 1.2:1 all the way from dc to 1500 MHz. A somewhat similar device, the CTC TC250-50, is rated at 250 watts, but the maximum frequency rating for 1.2:1 vswr is only 1000 MHz.

reference

1. Courtney Hall, WA5SNZ, "How to Solve Transistor Heatsink Problems," *ham radio*, January, 1974, page 46.

ham radio

the hand-held electronic calculator: solving problems

This article is the second in a series on the hand-held electronic calculator. The first article[1] described the operating principles of the four-function machine, gave examples of how to use it in handling basic arithmetic operations, and touched on the operations that can be performed with the more expensive and elaborate machines, including the use of constants, chain operations, and memories.

Despite its simplicity and low cost, the four-function calculator has as much potential calculating power as the more sophisticated machines, providing you're willing to accept slower operating speed and the occasional use of a scratch pad. The key to this power is in using approximations as a substitute for special calculator functions, using problem-organization techniques, and using a scratch pad as a substitute for calculator memory. With these tools, the simplest four-function calculator becomes usable for all radio calculations. The following text reviews many of the approximations that can be used and gives examples of problem solution. Clearly, the small calculator can be an instrument of great power.

calculator logic

As a first step it's desirable to spend a little time learning about the logic of the calculator. Sometimes the internal logic can be used to simplify a step; occasionally the logic will force a particular procedure. The major elements to understand are:

Rounding: Drop digits, round to nearest value, or hold internally.
Overflow, underflow: Signal used, calculation locked out or error possible.
Chain: How to interrupt; how to change from + − to ÷ X.
Constant: Switchable, automatic, chainable.
Sign: How to use; how to enter negative divide.
Memory: Self accumulate, overflow, sign.
Function keys: Determine logic (example: % is the same as = times 100 on divide, or as = divided by 100 on multiply).

These processes are best understood by working sample problems. For example, rounding can be checked by dividing 27 by 99 then multiplying by 2 (answer is 0.545454, rounds to 0.55).

problem organization

Most of the elements of problem organization concern arrangement of parentheses. For example, on the smaller calculators it's usually necessary to solve $Ax(B+C)$ by the steps $(B+C)xA$, the key strokes now being simple. Sometimes it's desirable to write the problem as $AB+AC$ to eliminate inner parentheses. Again, it's sometimes desirable to combine parentheses. For example, the form $(AxB) + (CxD)$ can't be solved directly on small calculators lacking memory but can be when rewritten in the form $\frac{AxB}{C} + D \: x \: C$. Quite often it's necessary to repeat a calculation several times using different values of the unknown. (An example would be the calculation of a set of half-wave filters for TVI elimination). For problems of any complexity it's worthwhile to lay out a tabular calculating form with the successive numbers across the top of the page and successive values of the unknown down the page. It's best to arrange unknowns *in order* to simplify a check for mistakes. This process can compensate for memory deficiency of the less-expensive calculators by providing a convenient place for recording intermediate results. The preparation of such a table has another benefit — it can be filed for a permanent record and as future reference.

accuracy and rounding off

The fact that even small calculators have 6 or 8 digits encourages the bad habit of regarding all digits as important. Sometimes they are, as when an expression of the form $(A-B)$ is encountered and A and B are nearly equal. But more often, only the first few digits have significance — the others may as well be zeros.

Many small calculators have provisions for rounding to two decimal places, which takes care of small numbers. For large numbers, or in the general case, calculator rounding is possible by temporarily converting to a small number then scaling back.

By R. P. Haviland, W4MB, 2100 South Nova Road, Box 45, Daytona Beach, Florida

If n is the number of digits to be rounded off, and m is the digit capacity of the calculator, divide first by $10^{(m+n)}$ then multiply by $10^{(m+n)}$. On most small calculators, m is the number of digits displayed; for rounding decimals, regard n as minus.

If the calculator truncates or cuts off the result instead of rounding, rounding can be obtained by first adding $0.5 \times n$ to the number then proceeding as above.

It's rare for radio calculations to require accuracies of more than a few percent, since this is the usual measuring accuracy For short calculations two significant figures are sufficient; three should take care of nearly all complex problems. An unnecessary number of digits gives a false impression of accuracy.

double precision

Despite the remarks above, there are times when the accuracy needed is greater than the number of digits provided by the calculator. Such problems can be handled by dividing the numbers into halves, performing a few operations, then combining the results – a process used in computer work called *double precision.* (Triple precision is possible but it's too complex to use).

Suppose a number *0123456789* is to be added to another number on a six-digit calculator. The first number *can be divided into two* parts, *01234* and *56789.* Denote the first part by A and the second by a. To add two such numbers Aa and Bb, compute the sums $C = A + B$ and $c = a + b$. Write the result as Cc. (There may be a carry from c to C or a borrow on subtract).

Example: Add 0123456789 to 1212121212

First number, Aa	$A = 01234$
	$a = 56789$
Second number, Bb	$B = 12121$
	$b = 21212$
Add $A + B$:	$C = A + B$
	$= 01234 + 12121$
	$= 13355$
Add $a + b$:	$c = a + b$
	$= 56789 + 21212$
	$c = 78001$

Write the result as Cc: *1335578001*
(sum of *0123456789* and *1212121212*).

If either number, say Aa, has more digits than the other number, say Bb, add zeros to the left of the smaller number. For example to add 0123456789 to 0123456, the numbers would be segmented as follows:

$$A = 01234 \qquad B = 00001 \qquad C = 01235$$
$$a = 56789 \qquad b = 23456 \qquad c = 80245$$
$Cc = 0123580245$, the sum

To multiply two such numbers Aa and Bb, first compute the four products $Cc = A \times B$, $Dd = A \times b$, $Ee = a \times B$ and $Ff = a \times b$. Compute the sums, $c + D + E = G$, and $d + e + F = H$. The final result is written as $CGHf$. (There may be a carry from f to H, from H to G, and from G to C). A little practice will help keep the partials straight.

Example: Multiply *1234* times *5678* (answer *7006652*)

$$\begin{aligned}
Aa &= 1234 \qquad A = 12 \qquad a = 34 \\
Bb &= 5678 \qquad B = 56 \qquad b = 78 \\
Cc &= A \times B = 12 \times 56 = 0672 \\
Dd &= A \times b = 12 \times 78 = 0936 \\
Ee &= a \times B = 34 \times 56 = 1904 \\
Ff &= a \times b = 34 \times 78 = 2652 \\
G &= c + D + E = 72 + 09 + 19 = 100 \\
H &= d + e + F = 36 + 04 + 26 = 66 \\
CGHF &= 06 \;\; 100 \;\; 66 \;\; 52 = 7006652
\end{aligned}$$

Note the carry when the second number, *100*, is combined with the first, *06.* An easy way to keep the carry straight is to write the numbers as

```
06
 100
    66
     52
--------
07006652
```

For division it's easiest to perform long division, which can be done by direct operations. Enter the leftmost digits of the dividend until the number in the display is greater than the divisor. Subtract the divisor and repeat until the remainder is less than the divisor, recording the number of subtractions as the first digit of the quotient. Multiply this remainder by *10*, which is the same as a *shift operation.* Add the next digit of the dividend and repeat the subtraction to obtain the next digit of the quotient. Repeat the shift-enter-subtract sequence until all digits have been entered. The remainder can be expressed as a fraction, or the division can be continued by entering zeros. If memory is available, the number of keystrokes needed is greatly reduced.

Example: Divide *987654321* by *12345* (answer: *80004.400243+*).

A. Enter *98765* (dividend) into calculator, then successively subtract *12345* (divisor) until remainder is less than the divisor:

```
  98765
- 12345    1st subtraction
  86420
- 12345    2nd subtraction
  74075
- 12345    3rd subtraction
  61730
- 12345    4th subtraction
  49385
- 12345    5th subtraction
  37040
- 12345    6th subtraction
  24695
- 12345    7th subtraction
  12350
- 12345    8th subtraction
      5
```

Since 8 subtractions were required, first (left-hand) digit of the quotient is *8*.

B. Multiply remainder by *10* and add next unused digit of the dividend: $(10 \times 5) + 4 = 54$

Since this number is still smaller than the divisor, no subtraction can be made so the second digit in the quotient is *0*

C. Multiply the remainder by *10* and add the next unused digit from the dividend: $(10 \times 54) + 3 = 543$

Since this number is still smaller than the divisor, no subtraction can be made so the second digit in the quotient is *0*.

D. Multiply the remainder by *10* and add the next unused digit from the dividend: $(10 \times 543) + 2 = 5432$

This number is still smaller than the divisor, no subtraction can be made, and the fourth digit in the quotient is *0*.

E. Multiply the remainder by *10* and add the next unused digit from the dividend: $(10 \times 5432) + 1 = 54321$

This remainder is larger than the divisor so successive subtractions can be made, as before:

54321	
− 12345	1st subtraction
41976	
− 12345	2nd subtraction
29631	
− 12345	3rd subtraction
17286	
− 12345	4th subtraction
4941	

Since four subtractions were required, the fifth digit in the quotient is *4*.

F. This same procedure can be used to as many decimal points as required, or the numbers (which are now small enough to be handled by most hand-held calculators), can be divided directly: $4941 \div 12345 = 0.400243$ Quotient is written from steps A through F:

$$ABCDEF = 80004.400243$$

table lookup

Because the small calculator doesn't have provisions for special functions or constants, these must be entered as needed. One way to do this is to keep a small set of tables with the calculator and look up values as needed. Small volumes of four-place tables of all common functions are often sold at school bookstores. Larger volumes are available, with greater accuracy, up to the 15-place tables issued by the U.S. Government. *Tables of Functions* by Jahnke and Emde is the standard volume for special functions.

Table lookup is also needed for typical constants and conversion factors. If several of these are needed often, it's convenient to make up a short table on *Dymo* tape or on a card and fasten it to the calculator case. Typical constants would be π, g, ft-lbs to ergs.

The second way of securing the functions needed is to compute them directly. Usually a full computation is too complex and time consuming to be practical, but many functions and constants can be generated by simple relationships that give approximately the correct value. Since most radio work doesn't require high accuracy, these approximations are extremely useful. The following material shows approximations and error for the commonly needed functions.

numerical values:

$\pi = 22/7$	error 0.04%
$e = 19/7$	error 0.15%
$1/M = 16/7$	error 0.73%

M is the natural log of 10, used to convert from Naperian (base e) to common or Briggsian logarithms (base 10).

log functions:

$e^x = 1 + x$	error less than 1% for $x < 0.15$
$e^x = 1 + x + \frac{x^2}{2}$	error less than 1.5% for $x < 0.50$

Convert e^x to 10^x by multiplying by $1/M$ (x positive):

$$log_e x = 2\left(\frac{x-1}{x+1}\right) + \frac{2}{3}\left(\frac{x-1}{x+1}\right)^3 \quad \text{error less than 1\% for } x < 2.5.$$

Convert to $log_{10} x$ by dividing by $1/M$.

trigonometric functions:

In the following, x is in radians or (degrees/57.296).

$sin\ x = x$	error less than 1% for $x < 14°$
$sin\ x = x - x^3/6$	error less than 1% for $x < 57°$
$cos\ x = 1$	error less than 1% for $x < 8°$
$cos\ x = 1 - x^2/2$	error less than 1% for $x < 38°$
$tan\ x = x$	error less than 1% for $x < 10°$
$tan\ x = x + x^3/3$	error less than 1% for $x < 30°$

For large values of x, or any value if desired, calculate from

$$sin\ (A \pm x) = sin\ A \pm x\ cos\ A$$
$$cos\ (A \pm x) = cos\ A \pm x\ sin\ A$$

The values of $sin\ A$, $cos\ A$ to each 10° can be kept on a small card or attached to the calculator case. The error will be less than 0.5%.

inverse trigonometric functions:

$sin^{-1}\ x = x$	error less than 1% for $x < 0.24$
$sin^{-1}\ x = x + x^3/6$	error less than 1% for $x < 0.58$

$\cos^{-1} x = \frac{\pi}{2} - \sin^{-1} x$	exact
$\tan^{-1} x = x$	error less than 1% for $x < 0.17$
$\tan^{-1} x = \pi/2 - 1/x$	error less than 1% for $x > 3.0$

It appears that all of the small calculators allow chained operation. This makes it easy to secure integer powers. Simply enter the number and press the times key the number of times of the power. Note that some calculators have automatic constant: for these it's best to press the times key once and the equals key the remaining number of times. If this isn't done, the calculator will retain the constant in a way that can cause errors. If the power is large it's faster to factor the power then take powers of powers. For example $x^{16} = \{[(x^2)^2]^2\}^2$, which saves eight key strokes.

Reciprocals, or any negative integer power, can be secured easily with any calculator having constant or memory, since $x^{-n} = 1/x^n$. Place the calculator in the constant mode, enter the value of x, press divide twice to secure a one, then press it n times for the desired value. For simpler calculators, the one must be entered, followed by a divide, then the number.

Non-integer powers are occasionally needed. Sometimes the unknown can be written as a difference, giving the approximations:

$$y^n = (1 \pm x)^n = 1 \pm nx$$

$$y^{1/n} = (1 \pm x)^{1/n} = 1 \pm x/n$$

$$1/y = \frac{1}{1 \pm x} = 1 \pm x$$

For values not close to unity use the approximation:

$$x^n = 1 + n \log_e x.$$

roots

Square roots are easily secured by a method of successive approximation developed by Newton. The value is obtained from:

$$\text{second estimate of root} = \left[\frac{\text{number}}{\text{first estimate}} + \text{first estimate}\right] \times 1/2$$

This is repeated until the digit requiring the desired accuracy doesn't change. This expression is one case of a general rule for any root, which can be written as

$$Z^r = \frac{\dfrac{Z}{A^{r-1}} + A(r-1)}{r} = A'$$

For example, for the cube root,

$$\frac{\dfrac{Z/A}{A} + A + A}{3} = \text{new estimate}$$

In these relationships, the number of operations is reduced if the first estimate is close to the true value.

A second way of securing integral roots is by trial. For example, if a cube root is needed, guess a number, cube it, and adjust the original guess. It's not difficult to secure two- or three-digit accuracy this way, but more accuracy becomes time consuming.

Non-integral roots can sometimes be solved as problems in powers. For example,

$$x^{0.8} = x^{(1-0.2)} = x \cdot x^{-1/5} = \frac{x}{\sqrt[5]{x}}$$

These roots may also be secured from

$$x^n = 1 + n \log_e x$$

hyperbolic functions

In transmission-line calculations, hyperbolic functions are useful. Some approximations for these are:

$$\sinh x = x + x^3/6$$
$$\cosh x = 1 + x^2/2$$
$$\tanh x = x - x^3/3$$
$$\text{arctanh}\, x = x + x^3/3$$

Note the close relationship to the trigonometric functions.

numerical integration

Integration, which corresponds to finding the area under a curve, is easily done using Simpson's rule. Its formula expression is:

$$\int_a^b F(x)\,dx = \frac{b-a}{3n}\,(y_0 + 4y_1 + 2y_2 + 4y_3 \ldots + y_n)$$

where $y_0, y_1, \ldots$ are the values of $F(x)$ at n equally-spaced intervals between a and b (n even).

To apply, plot the function as a curve, and divide the abscissa range into n equal spaces: ten is often a convenient number. Multiply each ordinate value by Simpson's multipliers, respectively, $1, 2, 4, 2, 4, \cdots 1$ for the first, second $\cdots$ last ordinate. Obtain the sum, and multiply this by the length of each interval divided by three.

concluding notes

These are by no means all the techniques known and used, but they should be ample for all radio work, and much more. The bibliography gives some further data and techniques. Particular attention should be given to the series obtained by expansion using Taylor's and MacLaurin's theorems. These are the basis of most of the approximations above.

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ham radio

automatic beeper

for station control

This simple one-transistor circuit improves communications efficiency on crowded channels

Amateurs who have heard radio conversations from the Apollo moonwalkers could not have helped but notice that the conversations ended with a beep sound, which indicated transmission termination. This beep sound is an "automatic over" signal. A communications system using such a signal saves time, words, and reduces misunderstanding. The beep-tone frequency is about 800 Hz, which seems to be the frequency most sensitive to the human ear. Proper installation of the beep system will also ensure that transmitter modulation is at a maximum level, just below flat topping.

The "automatic over" system was first used in the Yaesu FT2F 2-meter transceiver. The circuit described here is simpler than that used in the FT2F.

The automatic beep system works in such a way that, when releasing the PTT switch, a time delay occurs in the action of the changeover relay (or relays), which keeps the transmitter on the air just long enough to transmit a tone burst. This tone is generated in a simple one-transistor phase-shift oscillator, **fig. 1**. The circuit uses only a few components and is powered by the voltage present between the PTT terminal and ground in the receive mode. This voltage is commonly 12 Vdc but may be any voltage between 6 and 30 volts dc.

Turning on the transmitter grounds the PTT terminal so that the oscillator supply is shorted on transmit, thereby stopping the oscillator, which is on during receive. The sequence of events is:

1. Receiver on, beep-tone oscillator (BTO) on, relay opens.
2. PTT closes, transmitter on, BTO off, relay closes.
3. PTT opens, transmitter on, BTO on, relay delay effective.
4. Delay time finishes, relay opens, BTO on, receiver on.

construction notes

The 3.9k resistors and 0.01 μF capacitors should be 5% tolerance or closer. The 0.01 μF capacitors must be mylar, polycarbonate, polystyrene, or oil. *Do not* use ceramics. The 470 pF and 0.02 μF capacitors may be ceramic or other types. All resistors can be 1/2, 1/4 or 1/8 watt. The capacitors and resistors marked with an asterisk must be matched within 5%. If matched within 5%, a 20% variation from given values is satisfactory.

The transistor can be any small-signal silicon npn type with a gain of at least 300 at 1 mA. A low-gain transis-

By Earl Hornbostel, WA6URN, Republic Crystal Labs, P.O. Box 445, Greenhills Post Office, Rizal, Philippines

tor, mismatched resistors or capacitors, or poor-quality capacitors can cause the oscillator to start slowly or not start at all. Hundreds of different types of transistors are available that will work well in this circuit (i.e., 2N930).

There is nothing critical about layout so any method of wiring is acceptable. A small PC board can be made, assembly can be made on a perf board, or terminal lug strips can hold the parts. A convenient way of assembly is to make a small etched board from single-sided copper

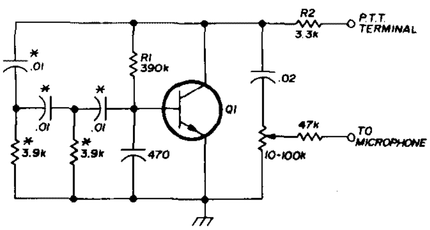


fig. 1. Circuit of the beeper oscillator. Parts marked with an asterisk must be matched within 5 percent. The transistor can be any small-signal silicon npn device with a gain of at least 300 at 1 mA.

clad stock, which is used as one side of a box on which the parts can be soldered on the *etched* side, leaving a continuous copper border. Do not drill any holes. The other five sides of the box can be made from the same material. One of these sides (not an end) can be used as a cover and should be slightly smaller all around, by the material thickness, so that it will fit inside the box formed by the other five pieces.

When the etched board has been wired and tested, the other five sides, with copper facing in, are then soldered together, taking care to line them up straight. Mount the microphone jack and dress the cable, leaving a little slack. Tape the cable inside the box to prevent strain on the soldered joints. The cover plate should have a single grounded wire soldered to it. With copper facing in, push the cover plate into the box about 1/8 inch (3mm), leaving an edge around which Duco cement can be applied. Don't use epoxy cement if you want to open the box later for servicing. This construction method results in a solid, shielded, insulated box that won't cause electrical noise if touched by a stray wire, nor will it be heavy.

installation

Only three connections are needed for this circuit: ground, microphone input, and the hot PTT terminal, all of which are in the microphone cable. The easiest way to install the unit is to use a duplicate microphone jack and plug and a small metal box. Mount the jack on one side of the box. On the opposite side drill a hole for a short piece of microphone cable and wire the duplicate microphone plug to the end of this cable. Mount the volume control to another face of the box. Connect the short length of microphone cable to the microphone jack. If a separate ground is used for the PTT circuit, (sometimes done to avoid switching transients that cause thumps during operation), be sure this ground is the same as that used for the oscillator. Also be sure the microphone shield does not short to the box or box wiring.

The last thing to do is to install a capacitor across the relay *first* controlled by the PTT switch. The capacitor, in the case of 12-volt relays, must be in the 300-1000 μF range with at least a 15-volt rating. This capacitor, which is shunted across the relay coil, will charge during transmit causing a delayed release when changing to receive, turning on the beeper. This action occurs because, while the relay is still holding, opening the PTT switch places a potential on the oscillator. The capacitor, in negative-ground systems, should have its negative terminal connected to the PTT-lead terminal of the relay.

No definite value for the capacitor can be given, since this depends on the coil resistance and the length of beep desired. The formula for calculation is:

$$T = RC \qquad (1)$$

where R = relay coil resistance (ohms)
C = capacitance (farads)
T = time (seconds)

The potentiometer should be set to fully modulate the transceiver without flat topping.

operation

The beeper will operate with any transceiver, a-m or fm, that has a PTT switch with a positive dc voltage of 6 to 20 volts on the PTT terminal during receive. This applies to most modern solid-state or hybrid sets that use a 12-Vdc relay for changeover. Tube sets may or may not have 12-Vdc control relays, so check this before starting the project. Be sure that the control circuit, when open, can pass 1 or 2 mA into the oscillator. In other circuits, such as when bias or the cathode of a tube is controlled, first check the voltage on the PTT terminal in the receive mode. Place a 10k resistor across this terminal and check again. If the voltage across this resistor is not less than 6 volts nor more than 30 volts; and if the receiver continues to operate, the circuit will permit the beeper to work. If negative voltage with respect to ground is present, then use a pnp transistor in the beeper.

Older sets, such as the KWM-2 and the other vacuum-tube equipment, may have the PTT relay in the plate lead of a vacuum tube. The much higher resistance in this case will not only require a lower capacitance, but a higher voltage rating, for the shunt capacitor. Some newer two-meter solid-state transceivers don't use a relay for switching. Instead, diode or regulator-type switching is used. In these cases, it will be necessary to use a much higher value for R1 – say 2.7 megohms or higher – and a high-gain (at low current) transistor. Not more than 50 microamperes should pass through R2, so that switching is not affected. The transistor is a low-current, high-gain device usually sold for "low noise" preamplifier use.

When the beeper is first used it will be difficult to break the "over" habit, but soon quick exchanges will increase the pleasure of operating and more natural conversation will ensue.

ham radio

turn-off timer

for portable equipment

This circuit prolongs battery life if you forget to turn off battery-operated instruments

Small battery-operated instruments are nice: no power cords, no heat, instant warmup, and portability. However, if you're like me you'll forget once in a while and leave something turned on with the result that one or more batteries will have to be replaced. So when I converted my vtvm to battery operation,[1] I decided to add a battery-saver circuit to prevent this from happening. The circuit worked out so well I have since added it to two other small pieces of test equipment.

basic circuit

Looking through the literature and reflecting on some recent and past experience, I decided to use CMOS devices and drive the ICvm load directly. There are several good reasons for this choice. The energized CMOS digital integrated circuit in the quiescent state uses "no power," (typically 0.009 μW per device at 9 volts),[2] output impedance is fairly low, input impedance very high, and they are inexpensive.

Using RC timing in a driving circuit to a CMOS gate or in a regenerative one-shot circuit is simple but has a big drawback in this application: the big "knees" in the transfer function. This simply means that voltage from an output gate will drift considerably over an appreciable part of the timing cycle. It's therefore necessary to drive the input gate with a reasonably fast pulse to avoid drifting through the transfer. There are several different ways this can be done, but the 2N6028 programmable unijunction transistor was chosen for the job. This device is stable over an extremely wide range of operating conditions and can be made to operate with practically no battery current. It's also readily obtainable on the surplus market for about 65 cents. Using this little gem with a CD4001 CMOS quad NOR gate resulted in the basic circuit of **fig. 1**.

operation

Gates G1 and G2 of U1 form a latch circuit or switch, which is operated by the Q1 timer. When S1 is switched to *on*, C2 charges through R2, causing the latch to switch making point A low (ground) and point B high (+9 volts). The timing circuit and the load connected to B are then turned on. R4 and R5 set the operating point of Q1. C1 charges through R1 then discharges through Q1. The discharge pulse across R6 is applied to G2 input, which switches the latch to its opposite state, A high B low. This action turns off both the timing circuit and the load, leaving only U1 in the quiescent state across the battery. Because U1 (the entire package) draws typically only 0.001 μA in this condition, S1 can remain *on* without decreasing battery life any more than its shelf life.

By Rich Hardesty, W5OXD, 2700 N. Lindbergh – Box 17, St. Louis, Missouri 63114

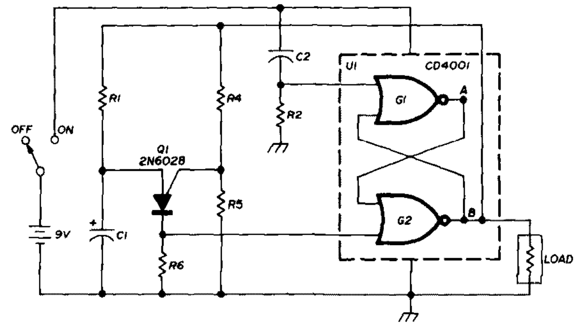


fig. 1. Basic ICvm timer circuit. Q1 is a programmable unijunction transistor, which drives a latch circuit composed of two gates of a CD4001 two-input quad NOR gate. With point A high and B low, both the timing circuit and load are turned off, leaving only U1 in the quiescent state across the battery. U1 current drain is only 0.001 μA in this condition.

complete timer circuit

Fig. 2 shows the complete turn-off timer for use with the ICvm. Resistor R6 of fig. 1 has been replaced with the primary of T1, a homebrew pulse transformer, the secondary of which is used with the negative supply to the meter. When the timing discharge pulse occurs, it will appear across both windings of T1, switching both latches at the same time to turn off the meter. With the resistor values shown in the Q1 circuit the timer current is maintained below 1 μA while on.

Diodes CR1, CR2 have been added to provide discharge paths for C1, C2 in the *off* position of S1 while maintaining timing-circuit isolation in the *on* position; a direct connection to S1 (negative supply *off*) forms the discharge path for C3. This allows the circuit to be retriggered at any point in the timing cycle by merely switching S1 *off* then *on* again. Note also that the spare gates in each of U1 and U2 are used to drive the meter circuit. This is done for two reasons – they are there, and paralleling outputs reduces the output impedance and drive losses. The small amount of battery power lost in the output of the drive gates when they are on is not power that can be usefully saved by operating the ICvm directly with the batteries. The output voltages to the meter will be slightly lower than that of the batteries, but the meter is calibrated at whatever voltages are supplied to it. If you want to adapt the timer to a single 9-volt operated device, the positive battery or upper portion of fig. 2 may be used by itself. In this case replace T1 primary with a 10k resistor.

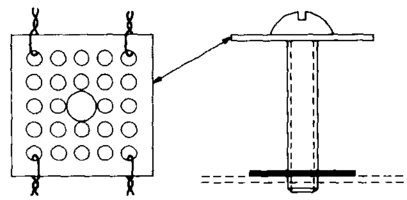


fig.3. Details of pulse-transformer core, which is a piece of 0.6 x 0.6-inch (15x15mm) perf board with 0.1-inch (2.5mm) hole spacing. Mounting screw should be nonmagnetic.

construction

T1 is made as follows: Construct the form as shown in fig. 3 using a 0.6 x 0.6 inch (15x15mm) piece of perf board (0.1-inch or 2.5mm hole spacing) for the top. Use pieces of solid hookup wire run through side holes and twisted to make four terminals. Make the center hole slightly smaller than an 8-32 (M4) screw thread and screw the top against the head of a 3/4-inch-long (19mm) brass or nylon screw. Then force a fiber washer onto the screw, leaving enough thread to mount the

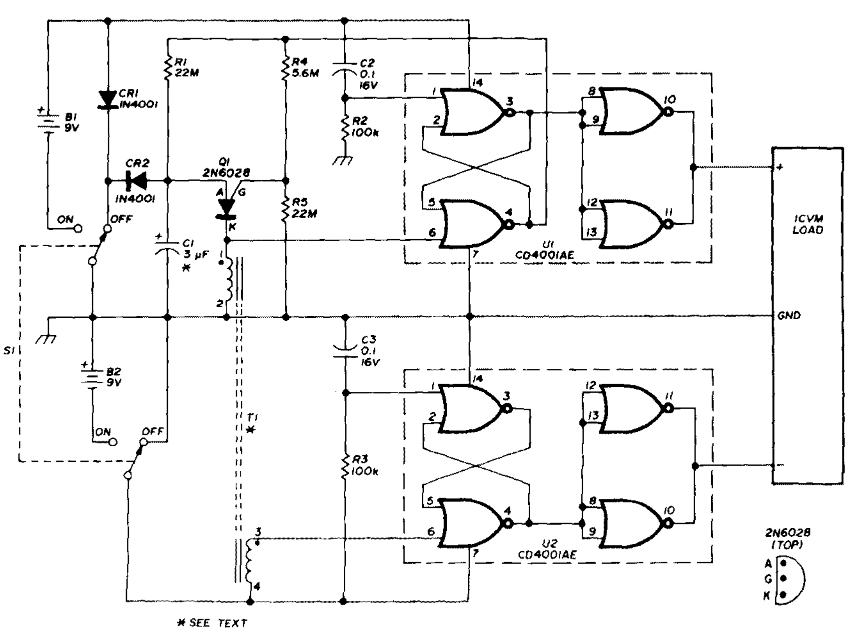


fig. 2. Complete circuit of the turn-off timer. All resistors are ¼ watt, 5% composition. Resistor R6 of fig. 1, which supplies the timing pulse to U1, has been replaced by the primary of T1, a homemade pulse transformer.

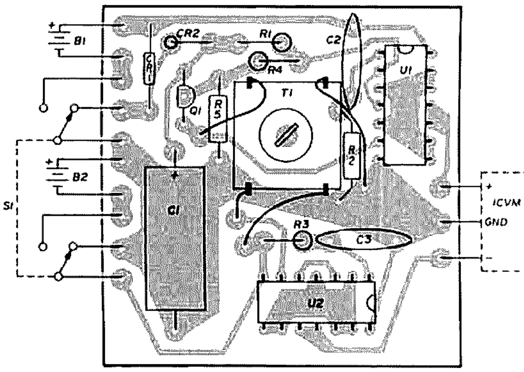


fig. 4. Suggested component layout for the ICvm timer. You can save space by using a lower-voltage capacitor for C1, which is a 3 μF, 50-volt unit.

finished transformer on the circuit board. Next, random wind on the form 600 turns of no. 36 to 40 (0.13 or 0.08mm) enameled wire, soldering the ends to the wire terminals on one side of the perf board top. Mark the start of the winding 1 and the end of winding 2. Repeat for the secondary, using the same number of turns wound in the same direction on top of the primary. Mark the start of winding 3 and the end of winding 4.

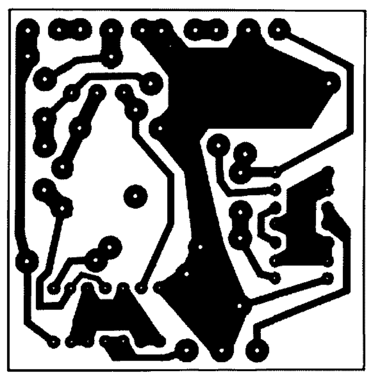


fig. 5. PC-board layout (foil side) of fig. 4.

Fig. 4 is a suggested component layout on a 2 x 2-inch (51x51mm) perf board or PC board. **Fig. 5** is a PC board layout (foil side) of **fig. 4.** C1 is a 3 μF, 50-volt capacitor. Obviously a lower-voltage type would take up less space. C1 should be a low-leakage type: polystyrene, polycarbonate, or mylar. With R1 of 22 megohms, I had about 3½ minutes-to-turn-off-time. The R1, C1 time constant may be adjusted to your preference.

references

1. Mike Kaufman, K6VCI, "How to Convert your VTVM to an IC Voltmeter," *ham radio*, December, 1974, page 42.
2. RCA Solid State 1974 Databook, Series SSD-203B, *COS/MOS Digital Integrated Circuits.*

ham radio

audio-frequency shift keyer

A stable, low-cost circuit designed around an 88-mH toroid

Many afsk circuits of varying complexity are available, but the circuit described here is the simplest and most satisfactory one I've found. Parts, excluding the relay,* are about $1.50. I don't know who the originator of the circuit is and claim no credit for it except to bring it to light. It was shown to me by WB6ETJ, who also uses it with complete satisfaction and who is equally in the dark as to its originator.

circuit

The circuit (fig. 1) uses one center-tapped 88-mH toroid tuned to the desired *space* frequency by suitable capacitors. These capacitors are paralleled by additional capacitors to tune the circuit to *mark* frequency. In the completed circuit, *mark* occurs when the relay is closed. For *space* frequency I used two capacitors totalling 0.0628 μF, then carefully removed turns from each side of the toroid center tap until the desired frequency was obtained. With care it's possible to obtain an accuracy of one or two hertz. The *mark* frequency may require three or four values of diminishing capacitance to reach the proper frequency, but with care equal accuracy is obtainable. If good quality mylar capacitors are used, the output is a perfect sine wave of equal amplitude.

A 2N404 pnp transistor is quite satisfactory as the oscillator; only 1.5 volts from a flashlight battery are required. The current drain is 100 microamperes, so battery life is very adequate. As I use my unit in a 12-volt system, the Variable Zener[1] set at 1.5 volts works very well. The output, variable from 20 millivolts to 0.08 volt, is controlled by a 10k potentiometer.

The relay coil is plugged directly into my 150-volt, 60-mA loop. When the loop current is turned on, the relay closes, and the afsk is on *mark*. Space frequency occurs when the relay is opened by the teleprinter keyboard.

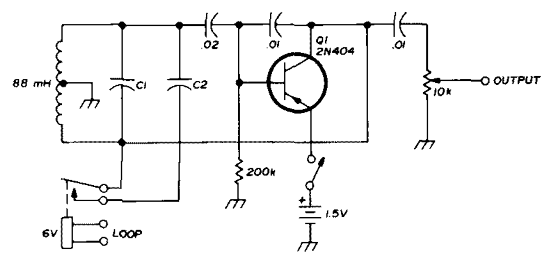


fig. 1. Circuit for simple, stable afsk. Capacitor C1 is tuned to 2295 Hz (approximately 0.0628 μF); C2 is tuned to 2125 Hz with relay closed (approximately 0.0156 μF). Q1 can be practically any pnp transistor. I.D. key may be the battery switch.

No special precautions are required for layout or construction. The unit works as well breadboarded as in a printed circuit I made to match my other gear. I've tried several types of 6-volt relays; all have worked satisfactorily. I've used this unit with an ST-6 and other terminal units. Since it plugs directly into the loop it doesn't interfere with other devices in the loop. All you have to do is plug the output into the transmitter audio input jack, and with a suitable output level you're in buisness.

reference

1. James McAlister, WA5EKA, "Low-Value Voltage Source," *ham radio*, November, 1971, page 66.

ham radio

*Available from Poly Paks, P.O. Box 942, Lynnfield, Massachusetts 01940. Order Clare reed sealed relay MR2MF-1006.

By John B. Dillon, M.D., KH6FMT, P.O. Box 758, Koloa, Kauai, Hawaii 96756

calibrating ac scales

on the vtvm, icvm and fet voltmeter

A simple circuit of emf sources for an indirect but accurate means of calibrating your test equipment for ac voltages

In the absence of a precision voltmeter, the dc ranges of a vtvm or its solid-state counterpart may be quite accurately calibrated by using a dc source of reliable known open-circuit voltage. The zinc flashlight cell commonly used for this purpose is not a very reliable source, but the readily available P-13 mercury cell is quite reliable.[1] The P-13 cell is well worth the cost of about $1.35 and possibly it could be stored for later use.*

calibration with the vom

Calibration of the vtvm ac scales to any degree of accuracy is usually not so simple. If you have access to a precision ac voltmeter you're fortunate, but this isn't the usual case. The vtvm ac scales are generally calibrated by the amateur against a vom. The resulting calibration can be, and often is, poor. Even the better voms usually have an accuracy on ac scales of only ±3 - 5% of full scale.

*It's possible that the P-13 mercury cell could be preserved for a long period by storing it in a refrigerator deep freeze section. I haven't tried this with mercury cells, but common zinc flashlight cells have been stored for a number of years at about zero F. (−18C) with little or no apparent deterioration.

When calibrating one instrument against another used as a standard, the errors of the two instruments are algebraically additive. It is for this reason that laboratories commonly use a standard instrument ten times as accurate as the one being calibrated. For example, to calibrate an instrument to ±1% accuracy, a standard instrument of ±0.1% accuracy is needed. In the worst case, let's assume the vtvm has an inherent accuracy of ±3%, is calibrated against a vom with an accuracy of ±5%, and the errors are accumulative. The calibrated meter would have an accuracy of not better than 8%.

calibration using ac line voltage

Probably a more common, but sometimes worse (and dangerous) method used by the amateur to calibrate the vtvm ac ranges is to use the ac mains and assume a nominal 117 volts. If the potential across the ac mains is indeed 117 volts calibration will be satisfactory, but how are we to know? The "nominal 117 volts" may very well be anything between 110 and 125 volts, and in some areas where electric power shortage is often critical, the emf at the mains could be 105 volts or less. Let's take a case where the vtvm is calibrated using the ac mains at an assumed 117 volts, but that voltage is actually 125 volts and the errors are accumulative. The voltage error of about 7% and meter error of 3% result in accuracy no better than 10%; not very impressive, to say the least.

calibration circuit

Fig. 1 is a simple circuit of emf sources for an indirect, but more accurate means of calibrating the vtvm ac scales. Ignoring for the moment the part to the right connected by dotted lines to Y1 and Y2, the circuit will be recognized as an ac supply — T1 with a half-wave rectifier, CR1, and a filter, C1. R1 plays no part in the filter and is only used to limit initial surge current through CR1 when power is first turned on; usually it may be omitted. There is no load across C1 except C1 leakage resistance and the backward resistance of CR1, both of which will normally be very high. Having negligible load, capacitor C1 will charge to the peak voltage of the ac emf across X1-X2 minus the forward voltage drop across CR1. T1 may be any step-down transformer. CR1 and CR2 are silicon diodes. R1, if used, may be anything between 22-100 ohms, ½ watt.

By K. F. Hager, W7KQ, Route 1, Box 186, Burton, Washington 98013

C1 and C2 are about 0.5 μF. (More about circuit parameters later).

procedure

Before calibrating the vtvm ac ranges, best possible calibration of the dc ranges should be obtained. The ac ranges are then calibrated as follows: Measure the dc voltage across Y1 and Y2 with the vtvm set for dc measurements, common lead to Y2, probe to Y1. Call this voltage E. Remove the probe from Y1, set the vtvm for ac measurements and connect the probe to X1. The rms potential across X1 and X2, E_{ac}, will be E *plus* the voltage drop across CR1, E_{cr}, multiplied by 0.707; that is: $E_{ac} = 0.707\,(E + E_{cr})$.

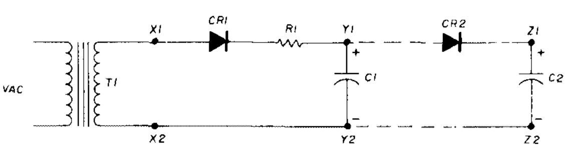


fig. 1. Circuit for calibrating the ac scales on a vtvm or icvm. The rms voltage, E_{ac}, across terminals X1, X2 is determined by E_{ac} = 0.7007 (E + E_{cr}), where E is the dc voltage measured across Y1, Y2 and E_{cr} is the voltage drop across CR1. Circuit at right of dashed lines is used to determine E_{cr}, as explained in text.

The ac calibration setup should now be arranged to show a meter reading E_{ac}. Example: Potential E is measured at 45 volts and E_{cr} has been determined as 0.3 volt. Therefore, 0.707 (45 + 0.3) = 32.0271 volts rms. Round this value to 32 volts. With the vtvm set for ac and probes across X1-X2, adjust the calibration so the meter reads 32 volts. Measurements should be performed several times to ensure that line voltage fluctuation doesn't introduce errors.

E_{cr} can't be measured directly with the vtvm but can be determined with a second like diode and a capacitor, CR2 and C2, connected to Y1 and Y2 as shown by the dashed lines in **fig. 1**. Do not attempt to measure the voltage drop by connecting the vtvm across Y1 and Z1. Shunting CR2 with the vtvm 11-megohm input resistance will cause the potential across Z1-Z2 to eventually reach the same potential as that across Y1-Y2. The speed at which this occurs, of course, depends on the capacitance of C2.

Measure the voltage at Y1-Y2 and the voltage at Z1-Z2. Note the difference in meter reading and let this difference be E_{cr}. CR1 and CR2 should be interchanged and the measurements repeated to determine any difference in voltage drop. Be sure to remove ac power and discharge the capacitors before starting a new set of measurements. In the calculation an ac sine wave is assumed; this will be found to be true where electrical power is from the U.S. and Canadian power grids. It may not be the case with a small, isolated power-generation and distribution system; in this event the waveform should be checked with an oscilloscope. Negligible load across Y1-Y2 is also assumed, and this is the case with a vtvm input resistance of 11 megohms.

circuit parameters

While nearly any available stepdown transformer may be used for T1, one that puts ac voltage across X1-X2, which permits ac and dc voltages to be read high on one scale of the vtvm, is preferable. If the vtvm has typical voltage scales of 0-1.5, 0-5, 0-15, etc., a small 6.3 V_{ac} filament transformer is satisfactory. The open-circuit voltage, which is essentially that across X1-X2, will probably be 7-8 volts or so; this permits reading ac and dc voltages well upscale on the 0-15 volt range of the vtvm. A better choice might be a transformer that places 32-35 volts across X1-X2. In this case, both readings would be taken high up on the vtvm 0-50 volt scale. A transformer placing 12-14 volts or so across X1-X2 would be a poorer choice, since the dc voltage would be read well down on the 0-50 volt scale. Connecting X1-X2 directly across the ac power mains is definitely *not* recommended. Not only would a situation similar to that just cited occur, but the practice is very hazardous.†

circuit constants

C1 and C2 may be almost any capacitors upward of about 0.5 μF. It is important that they have high leakage resistance. This requirement is easily met by tantalum or mylar-paper capacitors. Ordinary electrolytic capacitors may be used if the leakage resistance is high, although there appears little reason to use them. Capacitance is not at all critical. Capacitors between 0.22 and 80 μF have been used with no discernible difference in results. A difference in meter deflection just starts showing when the capacitance is reduced to 0.1 μF. Unnecessarily high capacitance is to be avoided, otherwise the voltage across C1 will change slowly when downward changes in ac line voltage occur. This lag makes calibration more time consuming, and possibly adds a little uncertainty in measurements. R1 may be omitted unless high capacitance is used.

Diodes CR1 and CR2 are International Resistor 170 or HEP 170, but they may be any good silicon diodes. Voltage drop across the HEP 170s, measured as described above, was 0.30 volt, with 45 volts across Y1-Y2. This drop remained constant with decreasing voltages across Y1-Y2 to 10 volts. With 5 volts across Y1-Y2, the CR voltage drop was 0.27 volt. Twelve HEP 170s were tested; 10 showed the same drop of 0.30 volt and two had just slightly less drop, with 45 volts across Y1-Y2.

In conclusion, there are a number of variables and therefore it's not possible to predict the accuracy to be expected after calibrating the vtvm ac ranges as described; however, it's likely to be better than that obtained by the usual methods of calibration.

†The vtvm common input is usually grounded to the chassis and metal case. If the common lead happened to be connected to the hot side of the ac line (a practically even chance), the full line voltage would be between the vtvm case and ground.

reference

1. Daniel A. Gomez-Ibanez, WB9ICI, "Calibrating a DC VTVM or FET VOM," *QST*, September, 1975, page 44.

ham radio

the MPC1000 — super regulator

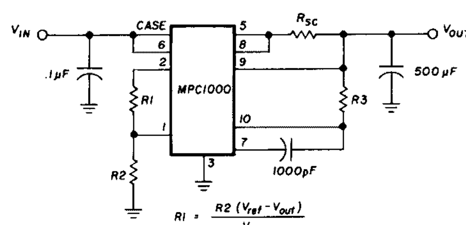
This versatile IC from Motorola provides up to 10 amps at any voltage to 35 volts — here are some typical applications

Just a few years ago a 5-volt, 10-amp, regulated power supply boasting 0.1 per cent regulation required many transistors, diodes and resistors. With the advent of power-regulator integrated circuits, the number of components has been reduced to less than a dozen. Now the Motorola MPC1000 reduces the complexity further as the need for an external pass transistor and associated current-limiting circuitry is eliminated.

While there are many applications where several 1-amp three-terminal regulators can be used, many applications simply require 10-amp capability from one regulator. For example, many solid-state power amplifiers require up to 10 amps at 28 volts with excellent voltage regulation to prevent generation of modulation products which result from changes in V_{cc}. Other uses include large digital projects such as counters, memories, and power supplies at a repeater site.

MPC1000 voltage regulator

The MPC1000 is a positive voltage regulator capable of providing up to 10 amps output current at any voltage to 35 volts. Certainly this device represents a further step in the development of voltage regulator ICs compared with the common 1-amp and 150-mA devices.



$$R1 = \frac{R2\,(V_{ref} - V_{out})}{V_{out}}$$

$$10k < R1 + R2 < 100k$$

$$R3 = \frac{R1 \cdot R2}{R1 + R2}$$

$$R_{sc} = \frac{0.66}{I_{sc}}$$

fig. 1. The MPC1000 as a positive voltage regulator for output voltages below the reference voltage, V_{ref} (7.15 ± 0.35 volts), including formulas for choosing the resistors that set the output voltage.

An additional feature of the MPC1000 is that it can be used in applications requiring negative voltages.

Since the MPC1000 can dissipate 100 watts, if you do require the full 10 amps you should choose a power transformer with a secondary voltage that doesn't ex-

By Gary L. Tater, W3HUC, 7925 Nottingham Way, Ellicott City, Maryland 21043

table 1. Electrical characteristics of the Motorola MPC1000 high-power voltage-regulator IC.

parameter	range or value
input voltage, V_{in}	+40 volts max
output voltage, V_{out}	+2 — +35 volts
voltage reference, V_{ref}	7.15±0.35 volts
output current, I_L	10 amps
internal power dissipation at 25C, P_d	100 watts
derate above 25C	0.667 W/°C.
quiescent current, I_q	5 mA
line regulation, V_{in} = 12-15 volts	0.1% V_{out}
line regulation, V_{in} = 12-40 volts	0.5% V_{out}
load regulation, I_L = 0.1-4 amp	0.1% V_{out}

ceed the output voltage from the IC by more than 10 volts. Table 1 lists the electrical characteristics needed to use the MPC1000.

The basic positive voltage regulator circuit for voltages below the reference voltage (7.15 ±0.35 volts) for the MPC1000 is shown in fig. 1. For output voltages above 7 volts, the circuit in fig. 2 should be used. Each includes equations for choosing the three resistors that set the output voltage.

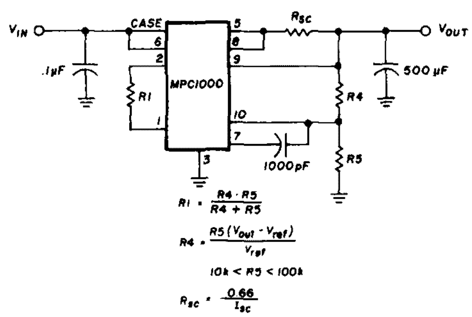


fig. 2. The MPC1000 as a positive voltage regulator for voltages above the reference voltage, V_{ref}.

Both circuits require a 0.1 µF capacitor at the IC input. If the input to the pass transistor (case) and the input to the rest of the regulator (pin 6) are separated, each would require a 0.1 µF capacitor. Such a situation would arise when you wish to operate the series pass transistor at a lower voltage than the rest of the IC to minimize heat dissipation within the device.

The current-limiting resistor, R_{sc}, typically runs from 0.66-0.066 ohms. If you have a resistance bridge, you can cut a length of wire to the exact resistance value by successively cutting off wire until that value is reached.

table 2. Data for computing the length of wire for the short-circuit sensing resistor, R_{sc}, in the regulator output circuit. Wire should be cut 50 per cent longer than lengths shown, then trimmed to obtain the required pass current.

copper wire size AWG (mm)	length ft (m)	resistance per 1000 ft (m) (ohms)
16 (1.3)	15.5 (4.7)	4.096 (1.25)
18 (1.0)	10.0 (3.0)	6.510 (1.98)
20 (0.8)	6.6 (2.0)	10.35 (3.15)
22 (0.6)	4.0 (1.2)	16.46 (5.0)

Table 2 lists information to approximate the length of wire needed. If you use this table, I recommend you cut the wire at least 50-percent longer than the calculated amount, then put a resistive load on the completed power supply. If your calculations were correct, the supply will immediately shut down. The short-circuit resistor can now be shortened by small increments until your current requirements are met. Although this procedure may require several cuts, it's the safest method to use if you don't have an instrument that measures low values of resistance.

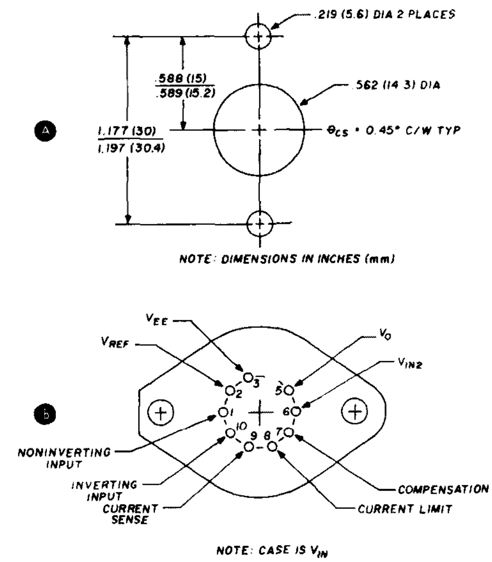


fig. 3. Mounting-hole pattern, A, and base diagram, B, of the MPC1000 voltage-regulator IC. A 9-pin tube socket makes a good mount — but observe pin-number sequence.

mounting

To realize the maximum ratings of this device, the MPC1000 should be mounted on a good heatsink. A simple heatsink mounting-hole pattern is shown in fig. 3A. I didn't have a 9/16-inch (14.5mm) punch, so I substituted a 5/8-inch (16mm) hole, which reduced the thermal path from the IC. A mica washer and socket are available from the manufacturer under the part number MK662.

If you want the regulator on the rear apron of a chassis to dissipate heat, a handy solution for making

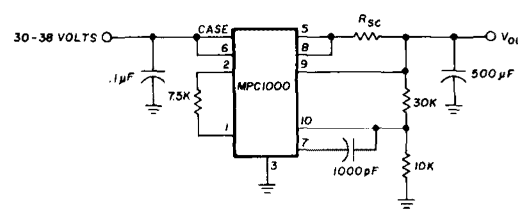


fig. 4. High-current, 28-volt regulator for high-power rf applications.

contact to the pins is to use a 9-pin miniature tube socket. Fig. 3B shows a bottom view of the pin connections. Note that the pins are *not* numbered sequentially.

practical circuits

I needed a 28-volt, 7-amp supply to power an fm Skyphone that I had converted from 459 to 432 MHz CW for Oscar 7, so I built the circuit in fig. 4 first. I used a 30-volt transformer with a bridge rectifier, which

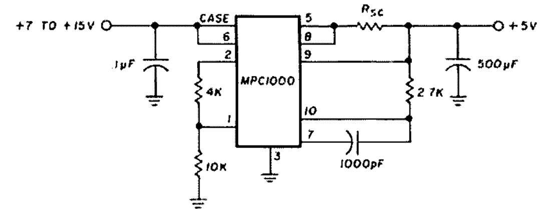


fig. 5. A 5-volt regulator suitable for high-current applications.

worked well when the short-circuit sensing resistor, R_{SC}, was adjusted properly. The initial shunt resistor was too long and caused the IC to shut down during key down. I removed two feet of wire from the homemade resistor, and the power regulator worked as advertised.

If you need several amps at 5 volts for a large TTL project, the circuit in fig. 5 should be useful. For a typical 8-digit frequency counter, which could require 3 or more amps, this one IC and its associated components could replace up to 4 three-terminal regulators.

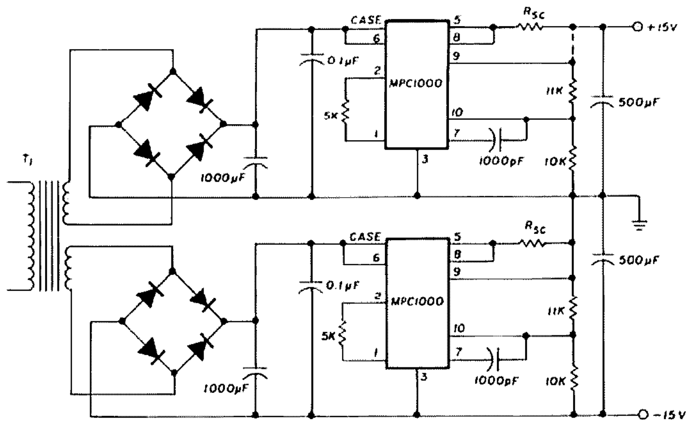


fig. 6. Regulator for providing +15 and −15 volts at 10 amps. T1 should supply 18-24 volts rms from each secondary winding.

The circuit in fig. 6 illustrates another feature of the MPC1000: two can be used to produce negative and positive supply voltages. The example in fig. 6 shows a high-current supply with a positive and negative 15-volt output.

Clearly, the MPC1000 is a versatile IC that can provide a quality high-current regulated power supply for your projects.

ham radio

how to clean printed-circuit boards

Here are some straight forward methods for cleaning PC boards after you have installed the components

There are still many amateurs who despite the efforts in obtaining parts, build their own equipment. The use of solid-state devices makes it mandatory that, at one time or another, the assemblies be done on printed-circuit boards. There has been a good deal of information published in the amateur magazines pertaining to board fabrication – from the original artwork through to the final piece. This short article presents information on the final step in the process and provides some answers to that age old question, "Is a clean board really necessary?"

All of us have noted that commercially manufactured equipment shows no globs of rosin or other soils (if they are there, don't buy it). Since it costs something to obtain that result, we must assume that there is a good electrical reason – as opposed to cosmetic -- for the added expense and there is. Further, there is an extensive military specification for board cleanliness.

To fully appreciate the problem and its solution let's talk a little about dirt. Since I have no intention of giving a chemistry lesson, suffice it to say that there are two kinds of soil that remain on circuit boards after you finish etching, drilling, inserting parts (also known as stuffing), and soldering. One is called polar or ionic and consists mainly of the residue from plating baths, etchants and, most important of all, residues of oil from your fingers. The latter is the most insidious. (As a point of information, it is an acknowledged industry fact that the residue from female fingers is different than that from male fingers). All the residues behave differently under varying climatic conditions. If your circuit board is to be used in a humid atmosphere -- and where is the place where there is no moisture in the air – the finger soil should be cleaned off with a solvent that will dissolve ionic material. I'll get into details in a moment because, as you will discover, we really need two different solvents for a perfect cleaning job.

The most noticeable residue found on PC boards and least potentially harmful is the hardened rosin residue of the flux. Rosin-based flux is a necessary evil used in all soldering operations. It is the best and safest remover of oxides of copper we have available and in order to make good solder joints the copper must be clean. Many claims have been made regarding so-called water soluble fluxes, which are highly active ammonium salts; I'm sure their use is advantageous in some instances, but I'm old fashioned and still prefer rosin-based systems. The residue from rosin-based fluxes is non-polar (non-ionic) and requires the use of non-polar or non-ionic solvents to dissolve it. Ionic residue can cause corrosion and shorting between conductors on a board because it can become conductive. On the other hand, rosin residues are insulators when dry and not decomposed; heat with humidity can cause rosin residues to decompose and become conductive.

How then can we insure that all our hard work will result in a working assembly over a long period of time under all conditions of environment? We clean the board with proper solvents using correct procedures. For the two types of soil there are two types of cleaner: polar and non-polar solvents. Water and alcohol are examples of polar solvents; chlorinated and flourinated hydrocarbons are examples of non-polar solvents. Trichloroethylene, trichloroethane, perchloroethylene and trichlorotrifluoroethane are some examples of non-polar solvents. Now that I have impressed you with a lot of fancy words, let's get down to practical cases. It is definitely advisable to clean off your boards, and there are several sources for some of the mouth-crackers listed above. It is certainly not difficult to find water although it is industry practice to buy de-ionized water to reduce contaminating residue contained in tap water. The alcohol is not the drinking kind but ethyl or propyl alcohol; 80% strength rubbing alcohol available in drug stores is suitable. Remember, the finger soil you want to remove is usually not visible, so just clean off the board with a brush dipped in the alcohol.

Obtaining and using a non-polar cleaner is not much more difficult. Trichloroethane, the recommended cleaning agent (also known as Trichloroethane 1,1,1;

By Budd Meyer, K2PMA, 6505 Yellowstone Boulevard, Forest Hills, New York 11375

chloroethene NU; VG), is available in electronic stores and from Miller-Stephenson* in spray cans. Not the least expensive way, but convenient. Freon TE is also available in spray cans. Another source, available in supermarkets in quart cans is *Afta,* a trichloroethylene-based household cleaner. Unfortunately trichloroethylene, as opposed to trichloroethane, can be hazardous if used improperly.

The methodology involved is to flood the area to be cleaned with clean solvent and scrub the area with a brush. I have found that the most effective method is to wet a local area, use a paste brush to scrub the area, then blow on the area or use a fan to assist in rapid evaporation. In industrial use the entire board is either immersed in the cleaner (except as noted below) or placed in a vapor degreaser. The degreaser consists of a large tank of solvent contained in a sump. Heating coils cause the volatile liquid to evaporate rapidly and become a "fog" just above the liquid. The boards are lowered into the fog and the condensate carrying the soil returns back to the sump. This allows the residues to eventually be separated from the fluid, which can be used many times after it has been cleaned up.

You must be cautioned that in recent years it has been discovered that chlorinated hydrocarbons wreak havoc with electrolytic capacitors. The most serious result is that the destruct phase doesn't occur until after many months or years of use. Avoid, with a passion, getting any of these solvents on or around the plugs (insulators) at the positive terminal of electrolytics. Alcohol won't cause any damage as far as is known — but neither will it clean up rosin residues. To emphasize how important this is, electrolytic capacitors that must be cleaned in trichlor now have an additional epoxy seal covering the positive end, and the manufacturers, having recognized the problem, are coming on line with new approaches to sealing the cans with new materials. Also be sure to keep trichlor away from polystyrene parts such as capacitors and trimmer capacitors. They will melt — literally! You should also be cautioned against using trichloroethylene in any place but a well ventilated area and to keep the amount of fluid coming into contact with skin to the barest minimum. If in doubt stick to trichloroethane.

Obviously, the ideal solution would be a mixture of polar and non-polar cleaners. These are available from solder manufacturers, usually under proprietary names, but unless you're willing to buy by the gallon, forget it. Localized cleaning is adequate for the small quantity boards amateurs have to make, and despite all the negative connotations noted above, it can be accomplished safely with care and common sense. Most of your cleaning will be on the copper side of the board away from the components. At the least, removing as much of the residues as possible results in a professional looking board, requires little effort, and will go a long way toward insuring reliability.

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*Miller-Stephenson Chemical Company, Inc., Danbury, Connecticut 06810.

repair bench

Joe Carr, K4IPV

troubleshooting transistor circuits

In this installment of repair bench I'll discuss troubleshooting of bipolar transistor circuits. The principal test equipment will be the vom or vtvm and, if you prefer, a transistor tester. First let's review some things about transistors that might affect servicing and trouble shooting.

preliminary considerations

In fig. 1A a stylized npn transistor is shown with its base-to-emitter junction connected across a power supply that causes it to be reverse biased. In this condition, charge carriers will be drawn away from the region at the junction, creating a relatively wide depletion zone. This allows only a very small current to flow across the junction. In fig. 1B the base-to-emitter bias voltage is changed so that the base-emitter junction is forward biased. Charge carriers are repelled by the battery polarity and are driven toward the junction barrier. Here they can combine with oppositely charged carriers from across the junction.

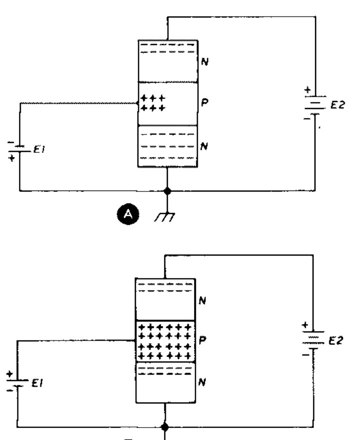

fig. 1. Relationship of charge-carrier propagation across the base-emitter junction in an npn transistor with reverse bias (A), and forward bias (B).

Figs. 2A and 2B show proper forward-bias-voltage relationships for npn and pnp transistors respectively. In an npn transistor circuit the base is positive with respect to the emitter by approximately 0.7 volt for silicon and 0.2 volt for germanium types. Pnp transistors have about the same values of voltage drop across the base-emitter junction, but it is of opposite polarity. On these transistors the base is more negative (or less positive) than the emitter. (Keep in mind that these polarities are relative quantities).

table 1. Junction voltages to be expected in a normally operating transistor amplifier. Values for silicon devices are shown, followed by those for germanium (in parentheses). Readings 1 and 3 were taken with the minus meter probe on the emitter; reading 2 was taken with the minus probe on the transistor base.

junction	npn	pnp
1. base-to-emmiter	+0.7 (0.2)	-0.7 (0.2)
2. collector-to-base	++	--
3. collector-to-emitter	++	--

A pnp transistor, as in the example of fig. 3, is still correctly biased despite the fact that the voltages on the elements are positive with respect to ground. Since the base is at 9.3 volts and the emitter at 10.0 volts, the base

will measure - 0.7 volt with respect to the emitter. **Table 1** gives voltage levels to be expected in a normally operating transistor amplifier stage. These voltages may not hold, however, in control circuits (such as squelch) or pulse circuits where the transistor may be reversed biased in one mode or another.

dc voltage checks

Any transistor stage has several dc voltages of interest to the troubleshooter. You want to measure the voltage drop across the emitter and collector load resistors and the base-emitter bias voltage. From these measurements you should be able to spot most faults. Of course, a circuit diagram that shows correct values would be of immense use, but it's not always necessary if you know what "ballpark" levels to expect.

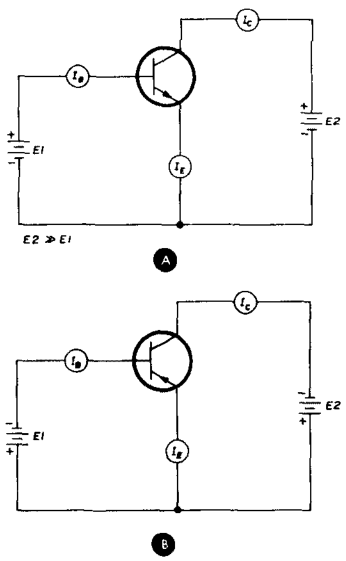

fig. 2. Proper forward-bias-voltage relationships for npn (A), and pnp (B), transistors. In (A) the base is normally positive with respect to emitter by about 0.7 volt for silicon devices and about 0.2 volt for germanium devices.

Perhaps the best method for learning the pattern of dc voltages that might be expected in the more common forms of transistor failure is the "case-history" approach. Assume that you have isolated the fault to a stage such as in **fig. 4** using signal tracing, signal injection or some other technique. Make the dc voltage measurements shown circled in **fig. 4**. Note that the emitter voltage is zero. Unless somebody has successfully repudiated Mr. Ohm, you're safe in assuming that the emitter current is also zero. You can also conclude that the collector current is either zero or at a very low value, because the collector voltage is close to the source voltage: a level about $V_{cc}/2$ would be normal. Measuring the base-to-emitter bias voltage, you find 10.7 volts instead of 0.7 volt. These symptoms usually point to an open base-to-emitter junction *in* the transistor. An ohmmeter or transistor checker will tell the tale if you're still in doubt.

Fig. 5 shows another common defect and its voltage relationships. The emitter voltage is about 0.27 volt, which results in a very low emitter current. Since you again have a collector voltage (17.4 volts) almost equal to V_{cc} (18 volts), you can say that no significant col-

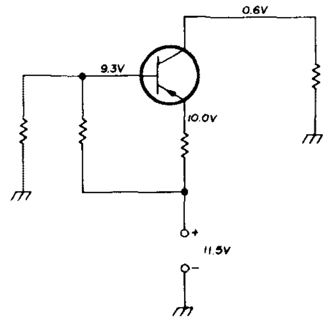

fig. 3. Example of a correctly-biased pnp transistor even though the voltages on the elements are positive with respect to ground. A vtvm measurement will show the base to be −0.7 volt with respect to the emitter.

lector-emitter current is flowing. The base-to-emitter voltage, however, is almost normal, so this junction is probably all right. In this case the collector-to-base junction is open.

A further common fault in transistor circuits is shown in fig. **6**. Measurements show a low collector voltage equal or a voltage very close to the emitter voltage. This transistor probably has a collector-to-emitter short circuit. If enough power has been dissipated, this effect may cause collector and emitter resistors to burnout. This condition almost always occurs in class-A audio amplifiers.

transistor testers

Once you've decided that a particular transistor is suspect, you might wish to make further tests using one of the many transistor checkers on the market. Simple instruments that claim to measure transistor beta are available at low cost, both ready built and in kit form. Be aware that the really inexpensive checkers might be

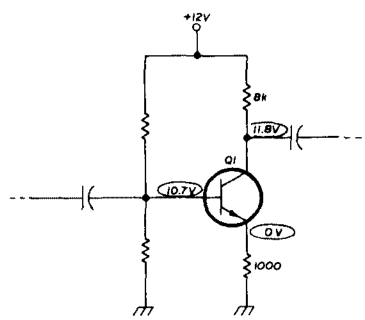

fig. 4. Example of the "case-history" approach to fault isolation in a transistor stage. Circled measurements indicate an open base-to-emitter junction.

too simple to give an accurate picture of the transistor's characteristics. The best transistor checker for ordinary service will provide some means for varying base current so you can determine whether the transistor base is able to control collector current.

An ohmmeter can be used as a transistor checker and can tell you quite a bit about the device under test. Certain precautions must be observed, however, as to the types of ohmmeters that are acceptable. Determine what type battery is used in the *ohms* section of your meter. If it's more than 1.5 volts, you may ruin as many transistors as you test! Some older instruments use battery voltages as high as 22.5 volts for the ohmmeter sections.

Alternatively, some modern fet voltmeters have an

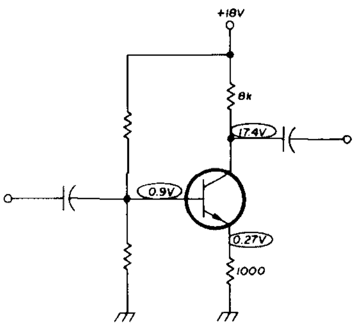

fig. 5. Another common defect in a transistor stage. Circled measurements indicate an open circuit in the collector-to-base junction.

ohmmeter source voltage that is too low for use as a "Transistor tester." This is a double-edged situation, because a desirable feature of such instruments is that they can be used to make in-circuit resistance measurements without removing the semiconductor devices. It is this very feature, however, that eliminates the instrument as a transistor tester. The dc voltage across the ohmmeter probes is too low to forward bias the pn junctions.

Test each junction separately by measuring its resistance twice (**fig. 7**). Measure the resistance the second time with the probes reversed from the direction used on the first try. Transistor junctions can be viewed as pn diodes and will not pass current in both directions under normal circumstances, so a very high resistance should occur in one direction and a much lower resistance in the opposite direction.

Normally, the reverse-forward ratio should be greater than 10:1. Check both base-to-emitter and collector-to-base junctions in this same manner. On power transistors use the RX1 scale, and on small transistors use the RX100 and RX1000 scales. If too low a resistance scale is used on those small transistors you may blow the junction.

Collector-to-emitter leakage may be checked in the same manner. Make two readings and take the higher one as the leakage resistance. The higher the better, and if it approaches your ohmmeter's idea of "infinity" so much the better. While making this test, you can also ascertain whether the base can control collector current. Connect the ohmmeter probes between collector and emitter. Next, short the base to the collector and note whether the resistance reading drops. If there is no response, reverse the probes and try again. If again there is no response, assume the transistor is dead.

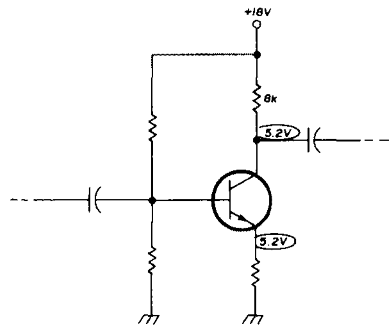

fig. 6. Circled measurements indicate a probable collector-to-emitter short circuit.

which stage is bad?

Signal tracing or signal injection techniques may provide the answer. However, a high-amplitude signal from a signal generator, or a transient generated when connection is made, may shock excite the defective transistor into normal operation. A far better technique would be to use dc analysis initially, then use one of the more traditional techniques only if the dc test fails.

Dc signal tracing requires only a vtvm or high-impedance vom. **Fig. 8** shows the transistor lineup in a

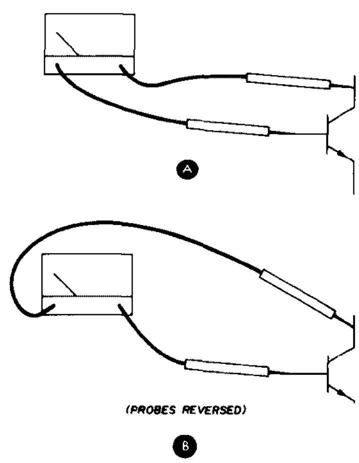

fig. 7. Using an ohmmeter as a transistor checker. Each junction resistance should be measured twice with the ohmmeter probes connected as in (A), then reversed, as in (B). If the transistor is good, the reverse-to-forward ratio should be greater than 10:1.

typical vhf fm receiver. This unit uses npn transistors, so ground the voltmeter minus probe, then use the positive probe to measure the voltage at each emitter in succession. Loss of emitter voltage indicates an open transistor or some defect that causes bias to be removed. The dc level tells you "this is the bad stage," and from there you can ascertain what's wrong. A higher-than-normal

emitter voltage, on the other hand, indicates a leaky or shorted transistor.

Rf amplifiers and some gain-controlled i-f amplifier stages may yield false negative results when this test is used because of agc action. If you can't disable the agc, try rocking the tuning dial back and forth while noting meter readings. The agc-controlled stage will show voltage variations as the dial is tuned across incoming stations. Although the agc can foul you up when troubleshooting, it can nevertheless be used to advantage. If the emitter voltage in this stage (or the collector voltage in some pnp rigs) varies as the dial is tuned across an active band, you can be pretty sure the defect is *not* between the antenna input terminals and the point where the i-f signal is sampled for the agc drive.

An S-meter can be used instead of a voltmeter as an overall check. Note whether the S-meter deflects normally as the receiver is tuned across the band. If it does then look elsewhere; if no S-meter deflection occurs, a problem exists within the agc loop.

A similar technique may be used to troubleshoot receivers that use pnp transistors, with modifications in procedure to account for the difference in transistor polarity. (In both cases assume that negative grounding is used – a fair assumption in most mobile equipment, but one possibly tinged with errors in some home equipment). In a pnp stage, connect the voltmeter positive probe to the B+ line and use the minus probe to measure the emitter resistor voltage drop (fig. 9). Most receivers use a fairly hefty electrolytic capacitor to decouple the B+ line; this component may often be used as a point of identification if no schematic is available. As in the case of npn stages, the voltage drop at the emitter resistor can give clues as to device malfunction.

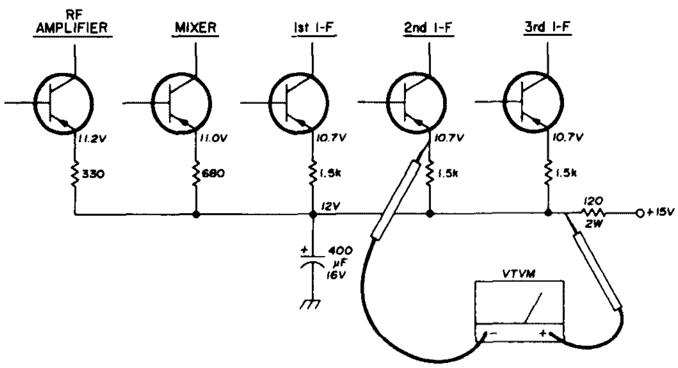

fig. 9. Using a vtvm for dc signal tracing in a vhf-fm receiver with pnp transistors. As in the circuit of fig. 8, the voltage measured at the emitter resistor will give clues as to device malfunction.

The oscillator stage can be checked using dc analysis techniques as an indicator of oscillation (but not of *oscillation* frequency). Connect a voltmeter across the emitter resistor, using whichever procedure is applicable to the transistor polarity. Tune the dial from one end of the band to the other. The emitter-resistor voltage drop will vary as the dial is tuned. This change will be greater on general coverage-receivers than on amateur-band-only models; but, even in the latter, some change will be noted. In crystal oscillators, sometimes a change in emitter-resistor voltage drop will occur when the crystal is removed from the circuit. In either case, the change identifies oscillation.

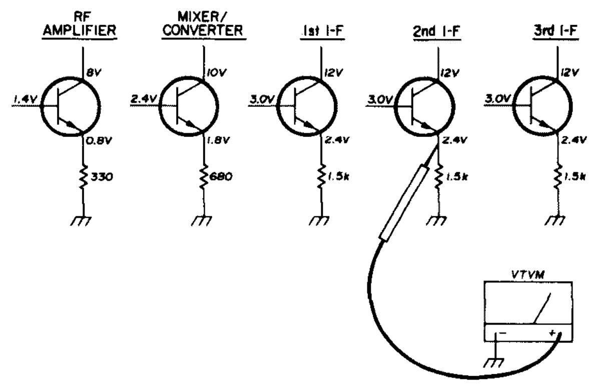

fig. 8. Dc signal tracing with a vtvm in a typical vhf-fm receiver using npn transistors. The probes should be connected as shown. Loss of emitter voltage indicates an open transistor or some defect that causes loss of bias.

caution note

Vacuum-tube test equipment is tolerant of some abuse, but transistorized equipment is not always so forgiving. Use of ungrounded ac-powered test equipment can easily generate both transient and steady-state voltage levels, which are quite capable of destroying transistors. Ground the cases of your test equipment. Two-wire power cords always identify ungrounded test equipment. For safety reasons it's wise to convert to three-wire power cords, in which the third wire is grounded to the chassis and cabinet. Of course, this only applies to equipment which is *not* ac/dc.

Grounded test equipment can create problems when troubleshooting, as in fig. **9**, where the grounded case might be connected to the voltmeter minus probe. In that case, you must use either a battery-operated voltmeter such as a vom or fetvm, or a voltmeter which, although ac powered, isolates both input probes from chassis. This, incidentally, seems to be the way many modern digital voltmeters are being made.

conclusion

The material presented here is intended for the average amateur who likes to service his own equipment. I've tried to cover the most likely problems that may be encountered. The procedures given should prove useful and will allow you to get back on the air as soon as possible without spending a lot of money for a repair bill.

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MICROPROCESSORS

interfacing a digital multimeter with an 8080-based microcomputer

This month we'll discuss the interfacing of an 8080-based microcomputer with a very versatile laboratory instrument, the Keithley model 160B digital multimeter with model 1602B digital output. We purchased this multimeter one year ago and found it to be an excellent example of what manufacturers can do to facilitate the interfacing of their instruments.

The Keithley model 160B is a general-purpose 3½-digit multimeter that can function as a dc voltmeter, dc ammeter, or ohmmeter. Twenty-six different ranges exist for the multimeter in its three modes of operation. The lowest range scales provide maximum readings of 1.999 mV, 19.99 nA, and 1.999 ohms. The 1.999 mV scale has an accuracy of ±0.1% of reading ±1 digit. Thus, a display reading of 1.000 mV will have an uncertainty of ±0.002 mV, or 2 μV. The highest possible readings associated with the three different modes of operation are 1200 volts, 1999 mA, and 1999 megohms, with the megohm reading accurate to only ±30%. This multimeter can be viewed as the digital complement of the ubiquitous multirange chart recorder.

The multimeter is basically a sophisticated analog-to-digital converter (ADC) that can handle most laboratory requirements for a digital data acquisition provided the data acquisition rate is no greater than one data point per second.* Switching between the 26 different ranges is performed manually. We would expect that, in the future, such switching will be performed by a built-in microprocessor operating under the control of an external computer.

By David G. Larsen, WB4HYJ, Peter R. Rony, and Jonathan Titus

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

The basic point of this month's column is the full interface circuit, shown in fig. 1, between the Keithley model 160B and a small development 8080-based microcomputer. The two OR gates and the SN74154 decoder generate the three different *device select pulses*[1] required to input data from the Keithley meter to the 8080 microcomputer. Note the $\overline{IN}$ input at pin 18 of the SN74154 decoder. This interface circuit takes advantage of the fact that all outputs from the 1602B digital output board are *open collector* and can be *bussed* together as in fig. 1. The noun, *bus*, can be defined as follows:[2]

> A path over which digital information is transferred, from any of several sources to any of several destinations. Only one transfer of information can take place at any one time. While such transfer is taking place, all other sources tied to the bus must be disabled.

*The settling time of the multimeter is about two seconds. Although five data conversions can be made per second, it may take about one second for the precision of a typical data point to reach 0.1% or 0.2%.

Notice how pins 16, 12, and 10 on the Model 160B are connected to the same input, D7, of the 8080 microcomputer. These three pins are said to be bussed together. Pins 35, 31, and 28 are bussed together to input D6; pins 17, 13, and 9 to input D5; pins 36, 32, and 27 to input D4; and so on. The eight inputs to the 8080, DØ through D7, comprise an eight-bit data bus over which information passes, one group at a time, from the Keithley multimeter to the 8080 microcomputer.

of data from the multimeter to the microcomputer over a set of eight data bus lines labeled DØ through D7.

A simple program that accomplishes the data transfer from the multimeter to the microcomputer is provided in table 1. The entire data acquisition and movement of data to registers C, D, and E occurs in 21 microseconds, a time that is fast when compared to the rate of five conversions per second by the multimeter. Clearly, considerable time is still available to the microcomputer

table 1. Microcomputer program that demonstrates the acquisition of 20 bits of data over the eight-bit data bus between the Keithley multimeter and the 8080 microcomputer shown in fig. 1.

10 memory address	instruction byte	mnemonic	clock cycles	description
000	333	IN 5	10	Generate device-select pulse that strokes the 10^0 and 10^1 digits into the accumulator
001	005	—	—	Device code for strobe inputs 1 and 2
002	117	MOV C,A	4	Move accumulator contents to register C
003	333	IN 4	10	Generate device-select pulse that strobes the 10^2 digit, the 10^3 bit, and the overload and polarity outputs into the accumulator
004	004	—	—	Device code for strobe inputs 3 and 4
005	127	MOV D,A	4	Move accumulator contents to register D
006	333	IN 3	10	Generate device-select pulse that strobes the Flag, $\overline{\text{Flag}}$, DP1, DP2, and DP3 outputs into the accumulator
007	003	—	—	Device code for strobe inputs 5 and 6
010	137	MOV E,A	4	Move accumulator contents to register E

At this point, 20 data bits are stored in registers C, D, and E. The microcomputer can now take this information and manipulate it in different ways. With the aid of the BCD digits and DP1, DP2, and DP3, it can determine the magnitude of the input decimal number. With the aid of the polarity input, the sign of the decimal number can be determined.

In the definition of a bus, it is indicated that only one transfer of information can take place at any one time. In fig. 1, this transfer is accomplished with the aid of the three sets of two strobe inputs. When a logic 0 is applied at strobes 1 and 2, the BCD codes corresponding to the 10^0 and 10^1 digits are transferred to the 8080 accumulator. The strobe signal for strobe inputs 1 and 2 is provided as a negative device-select pulse from channel 5 of the SN74154 decoder chip. In a similar manner, strobes 3 and 4 and also 5 and 6 permit the acquisition by the microcomputer of the remaining output data from the Keithley multimeter. In summary, three device-select pulses permit strobing twenty output bits

to manipulate the acquired data before new data is input into the accumulator.

Some additional explanation of fig. 1 is appropriate. Not shown in the figure are eight 4700-ohm resistors that are the *pull-up resistors* for the eight open-collector bus lines. One pull-up resistor is required for each of the eight data bus inputs. One end of the resistor is tied to +5 volts and the other end to the bus line. These resistors are not shown in the diagram because they can be added to the circuit board within the Keithley multimeter. The 8080 data bus normally employs an alternative bussing technique called *three-state bussing.* The interface circuit of fig. 1 represents a marriage of the two bussing techniques, open-collector and three-state. The 4700-ohm resistors do add a load to the data bus, but this does not prevent other devices from being tied to the bus provided each bus connection in the other devices can sink, in the logic 0 state, the additional 1 mA current produced by the 4700-ohm pull-up resistor.

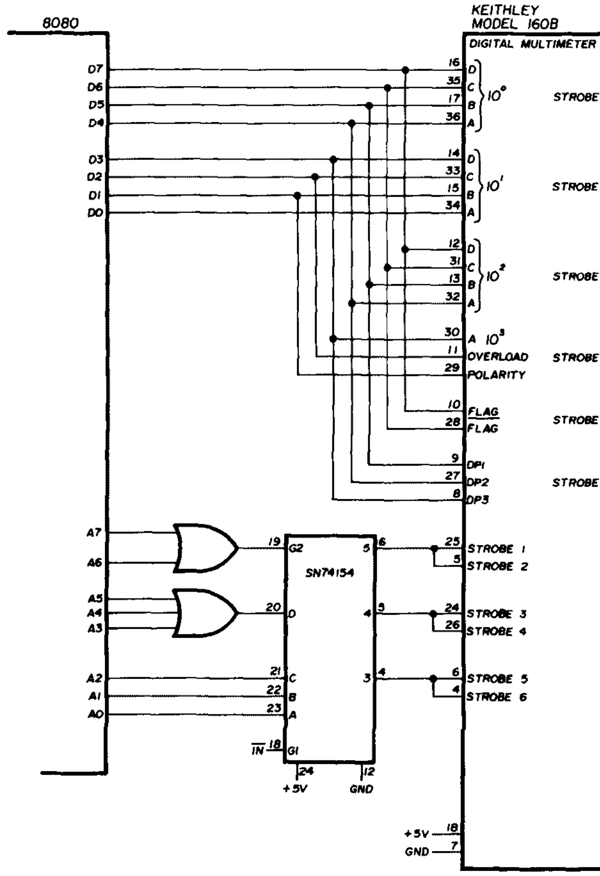

fig. 1. Interface circuit for a Keithley Model 160B multimeter and an 8080-based microcomputer.

At the beginning of this column, we stated that the Keithley multimeter is an example of what manufacturers can do to facilitate the interfacing of their instruments. In this case, what Keithley did was to provide open-collector outputs for all 20 output pins on the model 1602B digital output board. The added cost was small compared to the added value of the instrument. We expect future instruments to be microcomputer oriented in the sense that data bus outputs will be provided to permit direct interfacing of the instruments to microcomputers through simple wire interconnections. We hope these columns will encourage manufacturers to provide minicomputer- and microcomputer-oriented digital outputs and document such outputs as well as Keithley has done with their model 160B.

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KIT-A-MONTH

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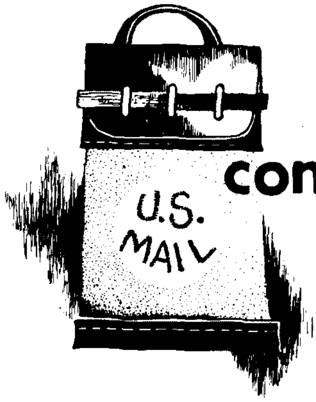
mits/2450 Alamo SE/Albuquerque, NM 87106 505-243-7821

Prices, specifications, and delivery subject to change.

mits/2450 Alamo SE/Albuquerque, NM 87106 505-243-7821



comments

fet-bipolar amplifiers

Dear HR:

I would like to comment on Irving Gottlieb's article, "A New Look at Solid-State Amplifiers," which appeared in the February, 1976, issue. The transconductance of all bipolar transistors is basically the same and equal to $1/r_e$ which explains why the spec sheets do not bother to mention it. The emitter resistance, r_e, has a value of 26 ohms at room temperature and 1 milliampere. This is because KT/q, the average energy of the current carriers, is 26 millivolts at this temperature. The bipolar Gm is thus 1/26 or 38,461 micromhos. The author by some roundabout calculations arrives at an erroneous figure of 300,000 micromhos for the bipolar Gm because he made the false supposition that the gain of the fet is 4 in his **fig. 1B**. The gain of the fet is the transconductance multiplied by the load impedance which is the input impedance of the bipolar transistor in this instance. The bipolar input impedance is 26 times beta. For the transistor the author considered, beta (grounded-emitter current gain) is 300 or the input impedance is 26 x 300 or 7800 ohms. The fet gain is then 4000 x 7800 divided by a million or 31.2. If the erroneous 4 is removed from the author's Gm_2 calculation and 31.2 substituted, the Gm_2 becomes 4000 x 300/31.2 or 38,461 micromhos.

The reader may conclude that the author's fet-bipolar combination produces greater gain than can be realized from a cascade pair of bipolars. The author seems to duck this comparison except to indulge in the debasing rhetoric of the bipolar's "current hungry base-emitter junction," etc. The gain of the author's pair is 1200 times the output impedance in kilohms. Two bipolars would supply a first stage gain of 38,461 x 7800 x 10^{-6} or 300 times while the second would have 38.46 times the kilohm output impedance. The total gain of the two bipolars is then 300 x 38.46 or 11,538 time the kilohm output Z. The gain superiority of the bipolar pair is thus 11,538/1200 or 9.615. We can then afford to lose 9.615 in a matching transformer to feed the "current hungry base" which means the impedance level can be 7800 x $(9.615)^2$ or 7800 x 92.45 which equals 721,096 ohms. Since the ordinary impedance levels are considerably less than this, the matching can be accomplished with less loss in gain and the bipolar combination will show a substantial superiority.

Since the author considered a bipolar type that is basically an audio transistor, I assume that his basic interest was at the lower frequencies. It should be pointed out, I believe, that as the frequency increases, vacuum tubes and fets can produce "current hungry inputs" of their own. This is because of negative feedback from common lead impedances and coupling capacitances. At 100 MHz, the input impedance of these majority-carrier devices may be no more than the bipolar's and thus with the necessity to match removed, the bipolar's superiority is even more telling.

J.A. Worcester
Worcester Electronics Laboratory
Frankfort, New York

To begin with, it is not *correct to say that the transconductance of all bipolar transistors is 1/26 miliohms, or 38,461 micromhos. If this were so, we would have long ago had a single "universal" transistor, rather than the thousands of types now extant. To bring this statement into a plausible ballpark, one would have to modify it as follows: "All bipolar transistors tend to develop the parameter of transconductance at the* rate *of 38,461 micromhos* per milliampere *of emitter current." Thus, we can expect to see evidence of my "alleged" 300,000 micromhos with the practically-reasonable emitter (and collector) current of about 7.5 milliamperes. Inasmuch as the silicon transistor tends to develop higher transconductance per milliampere than do germanium devices, one might obtain a transconductance of 300,000 micromhos with even less current, say in the vicinity of five or six milliamperes. (The current dependency of r_e, and therefore, transconductance, is clearly explained in the seventh edition of the* General-Electric Transistor Manual, *pages 45 and 46, under "Emitter Diffusion Resistance, r_e").*

Nowhere did I infer any intention to operate the bipolars at one milliampere. Hopefully, my bias networks will enable operation in the five to ten milliampere region. In the event the experimenter does not initially attain such operation, he can very easily bring it about by changing any one of the three emitter-base biasing resistors. The reason I mention this is that these inexpensive silicon-transistors often have beta tolerances as great as five to one. However, in a sample quantity of ten 2N3565s I found that seven of them had betas exceeding 250. So, I feel that my "nominal" beta of 300 is not atypical. Because of the current dependency of transconductance, you will find power transistors with transconductances on the order of 500 mhos!

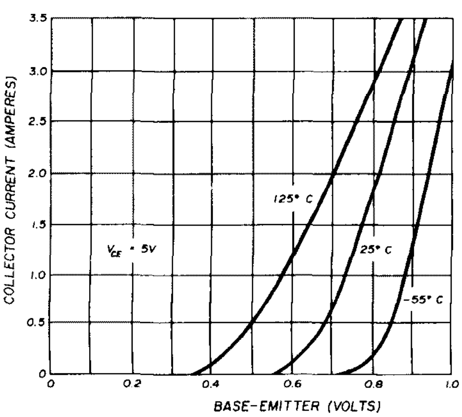
With regard to my phraseology in which I refer to "current-hungry base-emitter junctions", I view such a description as apropos when one compares the input circuits of bipolars with field-effect transistors. This is especially true when one recognizes that to obtain high transconductance from a bipolar transistor, it is necessary to operate at relatively high current levels – and this always tends to lower *the input im-*

pedance. It is no trick at all to operate fets with gate-source resistors in the tens or hundreds megohm range. This is not readily done with bipolars because of their "current-hungry base-emitter junctions."

Mr. Worcester introduces an interesting point in his discussion of transformer coupling. I acknowledge that transformers do *enable one to recover some of the loss in power gain that usually accompanies RC-coupled bipolar stages. Perhaps, somewhat arbitrarily, I decided to* exclude *the use of transformers in my article. The reason underlying this decision was that the average amateur does not have ready access to such transformers. Also, there can be a lot of headaches associated with coupling transformers. I wanted to provide* an *easy means of rolling one's own workhorse amplifiers.*

I agree with Mr. Worcester that at sufficiently high frequencies, even "infinite input-impedance" devices display dissipative, and other conductive effects. It is true that most of the applications I envisaged involved audio and low rf frequencies, say to several MHz. (Of course, even here, it behooves one to choose his devices carefully. I hope I have provided the readers with reasonable fets and bipolars insofar as concerns frequency capability).

I think Mr. Worcester will find that during the past five years or so, the major semiconductor vendors have included transconductance data in their specs. Below is a sample from Delco literature pertaining to their triple-diffused silicon power-transistor, the 2N5157. RCA, Motorola, and GE provide similar curves. Sometimes, however, the word "transconductance" is not used; the manufacturers merely depict collector current vs base-emitter volts with collector volts and temperature held constant.



DELCO 2N5157 TRANSCONDUCTANCE

Summarizing I contend that my numbers were reasonable, and that the fet would develop a voltage gain of about four because it "sees" approximately one-thousand ohms of bipolar input impedance. Moreover, the bipolar develops a beta of 300, together with a transconductance of 300,000. However, it operates between five and ten milliamperes, not *at one milliampere. I also concede that there are application areas where the fet-bipolar combo* does not *necessarily provide the best solution.*

Irving M. Gottlieb, W6HDM

old-time television

Dear **HR**:

Your well-researched article on television in the February, 1976, issue of *ham radio* was fascinating, although the results of K4TWJ's attempts to bring back 1925 TV to the amateur bands were a bit disconcerting. Apparently we haven't come as far on the road to deregulation as we thought.

Allow me one correction on your etymology: "Television" does not mean transmission of pictures over wires any more than "telescope" means seeing the stars by wires. *Tele-* is from the Greek and means "far off." Mr. Jenkins was entitled to call the new medium whatever he wanted, but he was no more correct in his naming than AT&T.

Joe Moell, WA6JFP
Fullerton, California

low-definition television association

Dear **HR**:

One of our members has sent me a copy of the article "50 Years Of Television," which appeared in the February, 1976, issue of *ham radio*. Clearly, American readers are not all up to date with developments outside the United States in the field of low-definition television (LDTV).

In 1971 I started LDTV experiments in collaboration with S. Kujawinski of Nottingham, unaware that Chris Long of Victoria, Australia, was working along similar lines at the same moment. A paragraph in the British magazine *Wireless World*, brought us into contact early in 1972. This magazine item revealed that Chris Long, in collaboration with Dan Van Elkan, VK3UI, had broadcast 48-line TV on the shortwaves on January 30, 1972, the first Australian LDTV broadcast since 1929. On the basis of this contact, I started a search for other correspondents and was so successful that on April 26, 1975, a formal LDTV society was formed. The Association is dedicated to reviving the techniques of the 1930s on an improved basis as a serious amateur activity.

Perhaps British experience of the LDTV period was less disillusioning than the briefer U.S. experience. This may be explained by the support given at the time by the giant BBC organization, the provision of an efficient carrier-borne synchronization system, powerful transmitting stations of huge range, daily programs, top-line nationally-known entertainers, and perhaps even the choice of a vertical scanning system. Certainly, to judge by the views expressed in the *ham radio* article, the cessation of the 30-line BBC-Baird system in late 1935, after six years, caused more consternation and anger in the U.K. than the corresponding termination in the U.S.A.

Although I can vaguely remember the old system as a boy, most of our members have only read about it in books, and our younger members (still at school) view it as the *very latest thing* with none of those "old-fashioned" cathode-ray tubes to worry about. The nostalgia mentioned in your article is almost totally absent here (in Europe and Australia). Our system is regarded as highly practical, relevant to present needs, and with synchronous sound, a valuable communication medium.

Members work on any standards they prefer, but for communication purposes the "Nottingham" standards have become widely accepted as a judicious compromise between the requirements of bandwidth and pictorial detail in the context of what can be achieved by reasonably skilled "kitchen table" engineers. Significantly, many of our members are ex-sstv enthusiasts or experimenters with a foot in each camp. The low cost of LDTV equipment and the "live" nature of the images are a tremendous attraction. No broadcasts

have yet been made in the U.K. (most members preferring closed-circuit experiments or "teletape" correspondence through the post, the extra stereo track providing the sound) but it should not be long before the once-familiar *ululation* is again on the air, if only on vhf.

For the information of your readers, the preferred standards of the LDTV Association are as follows: **Lines**, 32, vertical, with upward spot movement; **Frame speed**, 12½ per second; **Aspect ratio**; 3 vertical by 2 horizontal; **Line pulses**, blacker than black, 5 to 10% of line length (no frame synchronization pulses are employed, except for special purposes, e.g., closed-circuit usage); **Tape standards**, speed, 19 cm (7½ inches) per second; **Sense**, no signal equals black; **Tracks**, numbers 1 and 4, video, numbers 2 and 3, sound.

The LDTV Association is still welcoming new members and hopes to have representation in most European countries by the end of 1976.

D.B. Pitt
Chairman, LDTV Association
Nottingham, England

oscar communicators use excessive power

Dear HR:

After many nights and mornings of hearing Oscar 6 shut down because of overload and having myself "sucked out" of both Oscar 6 and 7 on numerous occasions, I feel that I must write to you in hopes that you will help me to make my thoughts known.

I am getting disgusted with the amateurs who completly disregard AMSAT's recommended power levels to hog the satellites for their own use or prevent others like myself, who try to stay close to the suggested power levels, from using the satellites. Face it, guys, you have no more right to use the satellites than anyone else. I imagine that you really do not care about how all that extra ERP affects the satellites' translators and power systems. One day you may find to your amazement that one of the birds does not work any longer.

The fellows in the Amsat organization have done all of us a great favor in getting the birds into orbit in the first place. Let's not blow a good thing. If I were Amsat and saw how the satellites are being abused, I would think twice about orbiting another Oscar, especially if it were as tremendously advanced as the proposed Phase III satellite.

So please, satellite users, refrain from using excessive ERP; it is not needed. If you are trying to access the birds with around 150 watts ERP and do not hear your downlink signal, you need to do one or more of the following:

1. Determine if you are really transmitting within the translator's receiver passband.

2. Find out just where your transmitter signal would appear on your receiver tuning dial.

3. Check to see that your antenna is pointing in the right direction.

4. Improve your receiver setup.

Notice that none of the above items includes increasing your effective radiated power.

Watson R. Gabriel, Jr., WB4EXW
Kernersville, North Carolina

DT-600 RTTY demodulator

Dear HR:

I always enjoy the projects and the informative articles published in *ham radio*. I just ordered a PC board from Data Technology Associates for the DT-600, and was very pleasantly surprised by the quality of the material and the amount of documentation enclosed with the board. They should receive a medal, And, believe it or not, the demodulator worked the first time I turned it on.

Rick Hall, K5GZR
Bellaire, Texas

two-meter frequency synthesizer

Dear HR:

I am writing to express appreciation for the extremely sensible two-meter synthesizer circuit published in the January, 1976, issue of *ham radio*. I hope to add this to my accumulating list of projects.

It is rather obvious that the state of the art dictates either a digital dial readout or outright synthesization, in the transmitting and receiving systems. Toying a bit with the basic control arrangement as a master control for transmitting, a couple of possibilities suggested themselves:

1. Providing a heterodyne crystal to present 140 to 144 MHz, feeding a decade divider, and thus affording the span from 14.0 to 14.4 MHz in 1 kHz increments.

2. Dividing again by two would present 7.0 to 7.2 MHz in 0.5 kHz increments, making a beautiful 40-meter CW unit.

3. Using the original 144 to 148 output, decade dividing to 14.4-14.8, and additive-heterodyning with 6.6 MHz would give 21.0-21.4 in 1 kHz steps.

4. Using the mode suggested in 2 above, and heterodyning the 7.0-7.2 with 5.2 MHz would give 1.8 to 2.0 MHz in 0.5 kHz steps.

5. Maintaining the original 144-148 output, subtractive mixing with 116 MHz would give 28.0-32.0 in 10 kHz steps for complete 10-meter coverage.

With the possibility of additional amateur bands becoming available after 1979, making just about everything but the "modifiable" Heath SB-104 broadband rig obsolete, I believe that a synthesizer useable as a master frequency control, whether it includes the necessary combinations for receiving or not, would be most welcome. My own training, skills and available time do not permit me to do much more than indulge in random periods of very wishful dreaming. I have a few bits and pieces of VHF engineering two-meter gear, and am convinced that they would blend perfectly with the basic synthesizer system to make a truly state-of-the-art transceiver for fm and CW operation.

Lee Clough, W5GQV
Texas State Technical Institute
Waco, Texas

sstv reporting system

Dear HR:

In October, 1966, sstv pioneer Copthorne MacDonald obtained permission to experiment with sstv on amateur radio and on Navy MARS. Since that time considerable effort has gone into technical improvement but not much

progress has been made in instituting a suitable reporting system. One of the projects of the Navy Marine Corps MARS SSTV Specialty Network, which was established in 1972, has been the development of a practical reporting system for general use.

The time-tested R-S-T system was a logical place to start. With readability (R), signal strength (S) and video quality (V) we came up with R-S-V. Several variations of designating the quality of video were tested, commencing with nine gradations. These were soon discarded as too cumbersome. Ken Wood, Jr., K6IIS and NNN0QXJ, eliminated all of the extraneous data and suggested a system that contains the key information you need 95% of the time.

This system has been in use for about two years with very good results on the Navy Marine Corps MARS SSTV Specialty Network which meets at 2100Z on Saturday and Sundays on 13975.5 kHz. It permits many reports to be made quickly and if you are so inclined you can make a chart of the reports and have a record of two-way propagation between a large number of stations scattered over a large geographical area.

The video reporting system varies from conventional R and S in that picture quality is not progressive. It is possible for a V3 picture to be better than a V4 picture. The system is as follows:

V5 Closed circuit quality pictures
V4 Good pictures with multipath
V3 Good pictures with interference
V2 Readable pictures with multipath and interference
V1 Mostly unreadable-loses sync-pictures interrupted.

This system is good for reporting radio and video reception on QSL cards, on voice or by video pictures and is particularily valuable for nets or contests where time is of the essence.

Thomas F. Pollock, WB6ZYE
Coordinator
U.S. NAVMARCORPS MARS
SSTV Specialty Network

European vhf-fm repeaters

Dear HR:

In *presstop* last year *HR report* editor W9JUV gave advice to those hams going to Europe to bring their portable two-meter equipment along with the appropriate crystals for repeater use. As we are using totally different channels for repeaters in Region I, compared to Region II, it might be interesting for your readers to know the correct channel numbers and the input and output frequencies for the European repeaters. Below is a short list which is self explanatory. In addition I can tell you that the international (within Region I) mobile/portable calling frequency is 145.500 MHz and the international mobile/portable traffic frequency is 145.550 MHz. In addition, there are twelve traffic frequencies on two meters (designated S21 through S33).

Gunnar Eriksson, SM4GL
Falun, Sweden

European vhf-fm repeater channels.

	in	*out*	*in*	*out*
R0	145,000	145,600	433,000	437,600
R1	145,025	145,625	433,025	437,625
R2	145,050	145,650	433,050	437,650
R3	145,075	145,675	433,075	437,675
R4	145,100	145,700	433,100	437,700
R5	145,125	145,725	433,125	437,725
R6	145,150	145,750	433,150	437,750
R7	145,175	145,775	433,175	437,775
R8	145,200	145,800	433,200	437,800
R9	145,225	145,825	433,225	437,825

European two-meter traffic channels.

S21	145,525	S28	145,700
S22	145,550	S29	145,725
S23	145,575	S30	145,750
S24	145,600	S31	145,775
S25	145,625	S32	145,800
S26	145,650	S33	145,825
S27	145,675		

microprocessor definitions

Dear HR:

I wish to note a peculiar inconsistency in the definitions given by Rony *et al* in their February, 1976, microprocessor article. On page 51, they correctly define "synchronous operation" as "operation of a system under the control of clock pulses." Yet, immediately above that, they define "synchronous computer" to mean "a digital computer in which all ordinary operations are controlled by equally-spaced signals from a master clock." In fact, as implied by the definition of "synchronous operation," there is no requirement for any synchronous machine to use *equally-spaced* clock pulses. Similarly, their definition of "synchronous states" . . . in which the performance of a sequence of operations is controlled by equally-spaced clock signals. . ." Again, while most (but certainly not all) synchronous machines use equally-spaced clock pulses, there is no switching-theoretical requirement for them to do so. (I believe that practical considerations almost invariably weigh on the side of using equally-spaced clock pulses).

Perhaps part of the explanation for this discrepancy lies in the fact that the correct definition of "synchronous operation" references the authors' *Bugbook III*, while the inaccurate definitions of "synchronous computer" and "synchronous" reference Graf's *Modern Dictionary of Electronics.* Rony *et al* bear partial responsibility for Graf's error by *printing it, especially when they* put it next to their own correct definition. This is not much ado about nothing: a misunderstanding of such fundamentals can foster much confusion; your readers depend on your accuracy, and you owe them no less.

E. Douglas Jensen
Sr. Principal Engineer
Computer Systems Research
Honeywell

Mr. Jensen is entirely correct in his observation: there is no switching-theoretical requirement for equally spaced clock pulses in synchronous machines. The definition given for a synchronous computer *applies to most existing computers, minicomputers, and microcomputers, and therefore should be a valid working definition for most individuals. The definition emphasizes the rule, rather than the exception. A more general definition would be as follows:*

Synchronous computer A digital computer in which all ordinary operations are controlled by clock pulses from a master clock.

Peter R. Rony
David G. Larsen
Jonathan A. Titus

the ham notebook

IC-230 mods: adding splinter channels

The ICOM IC-230 is a frequency-synthesized two-meter fm transceiver capable of operating on sixty-seven channels spaced 30 kHz apart, from 146.010 to 147.990 MHz. Four spare crystal sockets are provided to add 40 splinter channels. Each crystal adds ten channels spaced 30 kHz apart. The first of the four crystals are selected by placing the 146/147 switch in the 146 position and the 100 kHz switch in the A position. That is, 146.AX, where X is any of the ten positions of the 10-kHz switch.

The second crystal is selected with the B position of the 100-kHz switch; i.e., 146.BX. In a like manner, the last two crystals are selected with the 147.AX and 147.BX positions. The only difficulty is that the ten channels associated with each crystal are selected according to a strange coding scheme of the 10 kHz switch. The lowest-frequency channel is selected with the 1 position of the 10-kHz switch, and each successively higher-frequency channel is selected according to the following scheme: 1,4,7,0,3,6,9,2,5,8. Thus, the highest-frequency channels of the first splinter crystal would be selected by placing the controls to 146.A8.

You can select any number of crystals to give you any particular splinter channel. *I* have my favorite set of crystal frequencies to give complete coverage of all repeater splinters plus all direct splinter channels. If you don't want to spring for all four crystals for complete coverage, you can select your own. The formula for calculating the *lowest* frequency channel to the ten channels provided by a crystal is

$$f_x = \frac{f_L - 21.965}{9}$$

where f_x is the crystal frequency (MHz) and f_L is the desired lowest channel frequency (MHz). For example, if you want ten additional channels beginning at 146.415 MHz (the lowest direct splinter channel frequency), the crystal frequency would be (416.415 − 21.965)/9 ≈ 13.82778 MHz. When you install this crystal you'll add ten channels, from 146.415 to 146.685 MHz.

table 1. IC-230 splinter channels.

frequency (MHz)	setting	use	frequency (MHz)	setting	use
146.415	146.A1	direct	147.015	147.A1	repeater outputs
146.445	146.A4		147.045	147.A4	
146.475	146.A7		147.075	147.A7	
146.505	146.AØ		147.105	147.AØ	
146.535	146.A3		147.135	147.A3	
146.565	146.A6		147.165	147.A6	
146.595	146.A9		147.195	147.A9	
146.625	146.A2	repeater outputs	147.225	147.A2	
146.655	146.A5		147.255	147.A5	
146.685	146.A8		147.285	147.A8	
146.715	146.B1		147.315	147.B1	
146.745	146.B4		147.345	147.B4	
146.775	146.B7		147.375	147.B7	
146.805	146.BØ		147.405	147.BØ	direct
146.835	146.B3		147.435	147.B3	
146.865	146.B6		147.465	147.B6	
146.895	146.B9		147.495	147.B9	
146.925	146.B2		147.525	147.B2	
146.955	146.B5		147.555	147.B5	
146.985	146.B8		147.585	147.B8	

If you wish to omit the direct splinters and receive only the repeater splinter frequencies, you'd start at 146.625 MHz and your crystal frequency would be 13.85111 MHz. This crystal would provide ten channels from 146.625 to 147.195 MHz. For full coverage I recommend the following four crystals:

crystal freq. (MHz)	crystal socket (LO Module)	coverage (MHz)	setting
13.82778	8	146.415-655	146.AX
13.86111	9	146.715-985	146.BX
13.89444	10	147.015-285	147.AX
13.92778	11	147.315-585	147.BX

A decoding chart (table 1) is provided for using these crystals. When ordering crystals, be sure to specify *0.0025%, 20* pF in HC-25/U holders.

automatic offset switching

When operating the IC-230 two-meter fm transceiver, it is necessary to change the A/B switch to offset the transmitter either 600 kHz below or above the receiver frequency. Since most all "600-kHz low" repeaters are 146 MHz machines and most all "600-kHz high" repeaters are 147 MHz

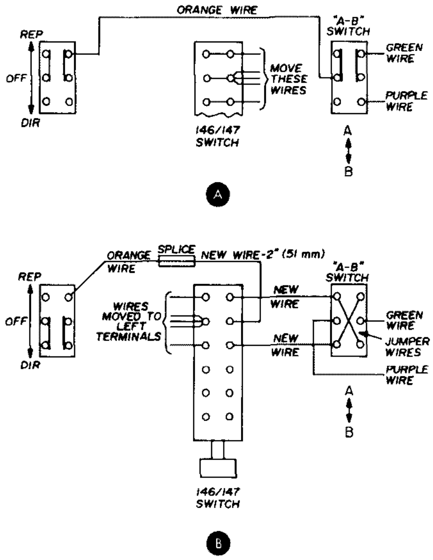


fig. 1. Existing wiring (A) and modifications for automatic offset switching (B) in the ICOM IC-230.

machines, it would be convenient to provide this switching automatically by the 146/147 switch. Fortunately, spare switch contacts are available that allow such a wiring modification to be made. The A/B switch then becomes a *normal-invert* switch, which operates as follows:

MHz switch position	A (normal)	B (invert)	transmitter offset (kHz)
146	X		−600
147	X		+600
146		X	+600
147		X	−600

After you make this mod you never need use the A/B switch unless you want to work "upside down" or work a 146.40/147.00 or 147.60/147.00 machine. Wiring details are shown in fig. 1. Here are the steps for making the changes:

1. Locate the 146/147 MHz switch and note that the rear set of terminals are wired in parallel. Remove the red wires, one at a time, and move them to the three left terminals thus making free the right set of three terminals. Clean the three right terminals.

2. Remove the A/B switch from its bracket. Remove the purple, green, and orange wires and clean the six terminals. Carefully install the jumper wires as shown in fig. 1B. Use AWG 26 (0.3mm) wire. Do not solder at this time.

3. Splice a 2-inch (51 mm) piece of wire to the orange wire that was removed from the A/B switch. Insulate the splice with heat-shrink tubing. Connect the free end to the terminal on the 146/147 switch (fig. 1B).

4. Connect two new no. 26 AWG (0.3 mm) wires from the two spare terminals of the 146/147 switch to the appropriate terminals of the A/B switch (fig. 1B). Solder the *outer* (corner) terminals only.

5. Reconnect the green and purple wires to the A/B switch center terminals as shown. Solder and check all connections. Reinstall the A/B switch.

Rod Cauvel, WA1OJX

tuning aid for the sightless

This audio tuning device is uncomplicated by connections other than those required to sample transmitter or exciter output at the coax transmission line. It uses no batteries and may be

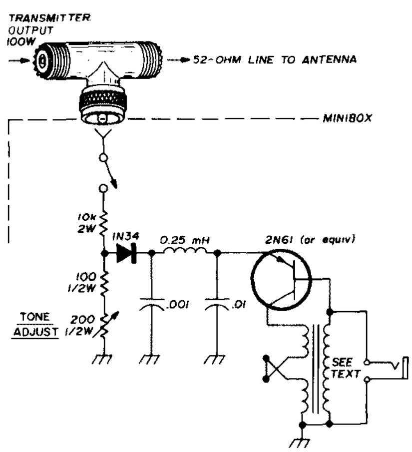


fig. 2. Audio tuning aid for determining maximum rf output power.

switched off after maximum rf output is determined. The circuit may also be used as a tone generator for monitoring CW keying; another possible use is a monitor for determining the condition of your transmission line (e.g., a short circuit between center conductor and shield).

The circuit is shown in fig. 2. A high-resistance voltage divider samples rf from the coax. The rf is rectified by a diode. This rectified voltage, which varies during transmitter tuning, is fed to a form of relaxation oscillator. Output varies in tone pitch as a function of voltage on the line: low voltage causes a high-pitch tone and high voltage causes a low-pitch tone, which indicates maximum transmitter or exciter output. The tone indication is similar to a dip in transmitter plate current.

The input voltage divider uses about 1 watt for 100 watts into a 50-ohm line. For higher power, the divider may be switched to higher values. For 1 kW, the input resistor should be about 100k, 2 watt. The schematic shows a voltage divider suitable for 100 watts output.

The 1N34 diode feeds about 2 volts to the transistor emitter, which draws less than 2 mA at maximum transmitter output. Any audio-type transistor may be used. If an npn device is used, diode connections should be reversed. The transformer is from a 5-watt transistor amplifier. Base and collector connections to the transformer can be reversed for needed feedback voltage. The color code on the transformer I used is: green to collector, red to base. The other connections in the collector circuit are: green/white to brown; brown/white and blue to ground. The transformer windings measure 22 ohms dc (high impedance); the other two, in series aiding, are each 4 ohm dc. Any transformer with similar resistance measurements should work. Audio output can be heard several feet from high-impedance phones.

D.H. Atkins, W6VX

HW-202 lamp replacement

To decrease the time and trouble of transceiver disassembly, try the idea shown in Fig. 3 the next time the dial light on your HW-202 fails. This lamp provides good illumination and has a prolonged life. Use caution when breaking glass on a burned-out bulb.

H.C. McDonald, W5UNF/6

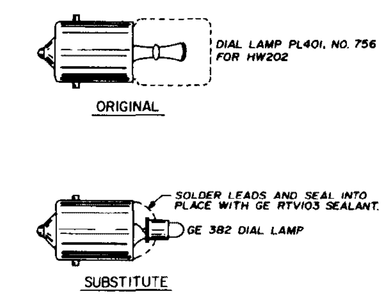
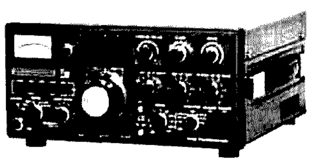


fig. 3. Simple lamp replacement for the Heathkit HW202.

NEW products

hf ssb transceiver

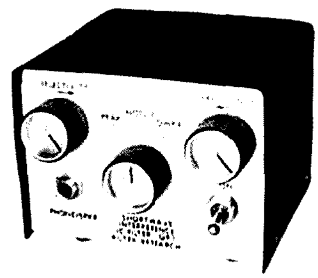


Trio-Kenwood has announced a new addition to its line of amateur radio equipment, the TS-820 "Pacesetter." Some features of this new rig are i-f shift, rf speech processing, digital readout (optional), digital hold (for the readout), rf monitor, monoscale vfo dial, PLL circuitry, rf negative feedback, RTTY, 160 through 10 meter coverage, vox controls, and rf attenuator. Probably the most unusual feature of the TS-820 is the use of a phase-lock filter in the local oscillator circuit, resulting in a very clean injection signal to the mixer. Speech processing is achieved at the rf level and is adjustable from zero to 50 dB through the use of a single, front-panel control.

A complete line of accessories will be available for the TS-820 including a plug-in digital kit, dc-dc converter (for mobile operation), 500 Hz CW filter, deluxe remote vfo, MC-50 microphone, two-meter transverter, and more.

The new TS-820 transceiver sells for $830.00 and is available from Trio-Kenwood Communications, Inc., 116 E. Alondra, Gardena, California 90248. Write for more information or use check-off on page 126.

all-mode active audio filter



The QF-1 "infinitely variable" filter from Autek Research is designed to combat shortwave interference with features never available to the amateur before. Peak, notch, and lowpass positions, all *continuously* adjustable over the range 250 to 2500 Hz, are provided for use in all communications modes.

The *lowpass* position rejects ssb and a-m hiss and splatter; the *notch* position rejects whistles, CW and carriers. The *peak* position provides superior CW reception with adjustable frequency for easy peaking, and bandwidth continuously variable from 50 Hz to flat. Skirts are steep and exceed 80 dB.

The notch tracks the *peak* frequency and is continuously variable from very wide to very narrow via the selectivity control. Notch depth is to 70 dB. The QF-1 drives a speaker or phones with room-filling volume (1 watt) and also includes a 117-Vac supply so there are no batteries to replace. It works with any receiver or transceiver by simply plugging it into the headphone jack. No impedance matching is necessary.

Selling for $54.65 postpaid, the QF-1 includes eight IC op amps plus power amplifier and metal cabinet. Order direct, or write for brochure to Autek Research, Box 5127E, Sherman Oaks, California 91403.

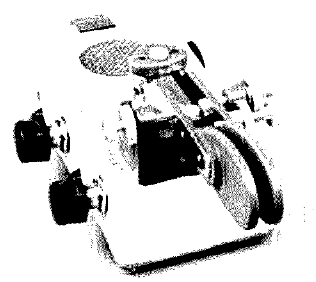
high power broadband isolator

Telewave, Inc. has recently introduced a new series of high power, broadband isolators which feature exceptionally low prices and require no tuning.

The T-1004 series operates in the 400 to 512 MHz bands and provides 30 dB of isolation over a 20 MHz bandwidth with only 0.3 dB insertion loss. The unit provides 60 dB isolation using two junctions and can be shipped standard single junction (30 dB), dual junction (60 dB), and triple junction (90 dB) in one module which eliminates cables and allows for extremely low insertion loss. All loads are removable and can be supplied in powers from 25 through 400 watts.

For more information contact Telewave, Inc., at 2166 Old Middlefield Way, Mountain View, California 94943 (415) 968-4400 or use *check-off* on page 126.

electronic keyer



Palomar Engineers has introduced a new IC keyer that takes less space on your operating table than the old semiautomatic mechanical key. The new keyer sends semiautomatic, full automatic, self completing, dot memory, iambic, or as a straight key. It has built-in sidetone oscillator and speaker, volume and speed controls, weight adjustment and battery holder. Any desired speed from 5 to 50 wpm can be selected while you send.

The IC keyer will key any transmitter, whether grid-block, cathode keyed or plate keyed, up to 500 volts and up to one ampere keyed current. Keying contacts are silver and withstand heavy surge currents and voltage spikes. The built-in paddle is fully adjustable for spacing and tension. A diecast metal case provides full rf shielding.

The clip-in 9-volt transistor battery will power the keyer for about 75 hours of normal operating, making the keyer ideal for portable operation. At the

home station, a lantern battery will last for about two years.

The keyer sells for $87.50 postpaid in the U.S. and Canada (California residents add sales tax). For more information write to Palomar Engineers, P.O. Box 455, Escondido, California 92025 or use check-off on page 126.

push-to-talk microphone

New from Astatic is the D104 *Silver Eagle* microphone with push-to-talk efficiency. The push-to-talk bar has been added to the D104 "grip-to-talk" desk stand for convenience. A slide lock clamp provides easy "no hands" transmission.

Factory wired for universal hook-up application, the *Silver Eagle* can be converted to electronic or relay operation. The microphone is wired with an open audio line on receive and comes with a coil cord with single-conductor shielded plus 4 unshielded. Switching requirements are determined by the proper hook-up at the cable plug end. All external parts, including the base, are chrome plated.

Information on the D104 microphone can be found by writing Astatic, Conneaut, Ohio 44030, or by using the check-off on page 126.

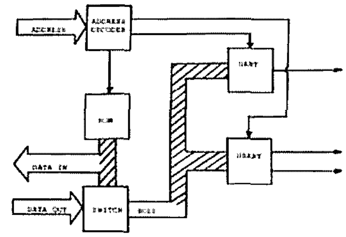
ARRL electronics data book

The new *ARRL Electronics Data Book* has been written for all technical levels from the beginner to the graduate engineer. Edited by Doug DeMaW, W1CER, it contains a compilation of essential data which is normally scattered among several reference books that an amateur or professional person would maintain in his library.

Among the many subjects treated in depth are RF Circuit Data, L, C and R Networks, Broad- and Narrow-Band Transformer Design, Modern Filter Design, Antennas and Feed Systems and a Catalog of Practical Solid-State Circuits. All chapters include pertinent simple equations with examples worked out to illustrate how the solutions are obtained. The section on transformer design deals mainly with toroidal broadband components of the conventional and transmission-line varieties. Schematic and pictorial diagrams are furnished for each transformer type. The chapter on modern filter design covers two- and three-pole Butterworth derivations for most of the frequencies of interest to amateurs. Tables of practical filter values are also included.

This book is an essential adjunct to the *Handbook* and other ARRL technical publications. Soft cover, 8¼ x 11 inches (21x28cm), 128 pages. $4.00 in the U.S.A. and Possessions, $4.50 elsewhere. Order your copy from Ham Radio Books, Greenville, New Hampshire 03048.

I/O board with rom

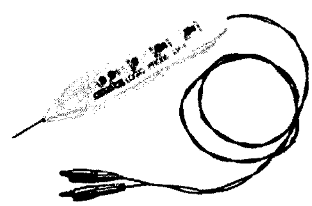


A new "Turnkey" I/O board is now available for 8080 microprocessors. The user is now able to turn on his computer and "go." All bootstraps, loads, dumps and edits are in a 512-byte ROM on the I/O board.

The board utilizes a UART to interface with a terminal and a USART to interface with one or two cassettes to make a complete computer operating system. Just type in what you require and the ROM executes the program.

The Turnkey I/O board is available in kit form for $140 or for $170 assembled and tested from National Multiplex Corporation, 3474 Rand Avenue, Box 288, South Plainfield, New Jersey 07080 or use the check-off on page 126 for more information.

logic probe



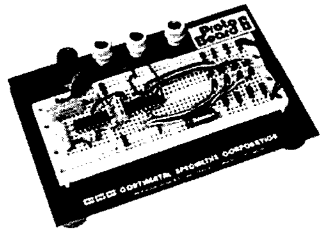
A low-cost, pocket-sized, multifunction test instrument known as the model LP-1 *Multi-Family Logic Probe* is available for digital applications from Continental Specialties Corporation. The LP-1 combines the functions of a pulse detector, pulse stretcher and memory circuit, allowing amateurs to get an instant picture of static and dynamic circuit conditions with most popular logic families. LP-1's ability to detect pulses as short as 50 nanoseconds, coupled with its stretching and latching ability, means that one-shot, low-rep-rate, narrow pulses — nearly impossible to see, even with a fast scope — are now easily detectable and visible.

The user simply connects the clip leads to the circuit's power supply, sets the logic family switch to the proper position (TTL/DTL or CMOS) establishing the correct logic level for the family under test and touches the probe tip to the circuit node. Two level detector LEDs plus a blinking pulse detector LED display signal activity at the node under test. At high frequencies, LP-1 will also indicate whether or not signals are symmetrical.

High input impedance on both DTL/TTL and CMOS modes virtually eliminates loading problems in the circuit under test, and input impedance is constant for both logic 1 and 0 states. The LP-1 sells for $44.95.

For more information, use the check-off on page 126 or contact Continental Specialties Corp., 44 Kendall Street, Box 1942, New Haven, Connecticut 06509.

solderless breadboard kit



Proto-Board® -6, a low-priced solderless breadboard kit, is available from Continental Specialties Corporation. This compact kit can be assembled in minutes and offers six 14-pin DIP IC capacity for basic breadboarding, testing and building applications.

The PB-6 includes one QT-47S solderless breadboarding socket, two QT-47B bus strips, four 5-way binding posts, a metal ground and base plate, rubber feet, all nuts, bolts, and screws, plus complete easy-assembly instructions.

The PB-6 lets the user test and build circuits without soldering or patch cords; all interconnections between components are made with common no. 22 AWG hook-up wire. Measuring 6 inches long by 4 inches wide (15cm x 10cm), the PB-6 sells for $15.95 from local distributors or direct from Continental Specialties Corporation, 44

Kendall Street, P.O. Box 1942, New Haven, Connecticut 06509; 351 California St., Box 7809, San Francisco, California 94104. For more information use check-off on page 126.

National TTL Data Book

A new TTL Data Handbook describing National Semiconductor's complete line of bipolar logic devices has just been published. The new handbook gives full specifications and electrical performance characteristics on standard 54/74 TTL, low-power 54L/74L, high-speed 54H/74H, ultrahigh-speed Schottky 54S/74S, low power Schottky 71LS/81LS, Series 9000 TTL, Series 10,000 ECL, and Series 930 DTL. An industry cross reference guide and a functional index are also included, as well as package outlines.

The TTL Data Handbook may be obtained by sending a $4 check (California residents add 6% sales tax) to the Marketing Services Department, National Semiconductor Corp., 2900 Semiconductor Drive, Santa Clara, California 95051.

concealable antenna mount

A steel antenna mount which stops radio antenna theft, trademarked *Tuk-A-Way*, has just been introduced. Sold as an accessory for citizens-band radio, mobile radio and car telephone users, *Tuk-A-Way* installs easily on the trunk lip of most car models, and provides complete antenna concealment inside the trunk when not in use. It accepts antennas designed for either roof or trunk mounting.

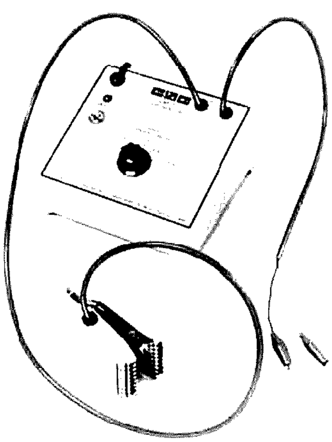
Tuk-A-Way offers three essential benefits for car radio and telephone users: the added protection against theft; the convenience which allows use of automatic car washes and covered parking facilities; and the elimination of

paint chipping and scratching or permanent holes in the roof or trunk by use of a clamp for installation.

For durability, the antenna mount is constructed of 12-gauge, cold-rolled steel, and is coated with zinc chromate. The hinge is of stainless steel and is spot-welded to the clamp to provide positive ground contact. In addition, the hinge is fitted with a stop which holds the antenna suspended and off the trunk floor while stored, allowing for short-range reception from inside the trunk.

Tuk-A-Way is available for $15.95 from Deep South Marketing Corp., 2828 Telephone Road, Houston, Texas 77023.

logic monitor



Logic Monitor 2, a digital test instrument incorporating a fully isolated power supply and selectable trigger threshold which matches the precise characteristics of the logic family under test, has just been introduced. The LM-2, which lists at $125.00, consists of two units; a connector/display unit which clips over the IC, and a power-supply module, which contains the precision reference power supply and logic family selector switch. Because it simultaneously displays 16 channels of information, it can show the user far more than an oscilloscope, and it's always automatically in synchronization.

LM-2's self-contained power supply means that there is no loading of a circuit under test, avoiding the problem of logic level shifts, false triggering and power-supply loading which sometimes occurs with other types of equipment.

In use, the LM-2 visually displays gate inputs rising and falling, passing pulses from circuit to circuit. Flip-flops may be seen changing states; decoders and encoders can be seen accepting and recoding information.

Further information on the LM-2 can be obtained from Continental Specialties Corp., 44 Kendall Street, Box 1942, New Haven, Connecticut 06509, or by using the check-off on page 126.

hand-held ssb transceiver

The MM-2C is a hand-held, single-channel ssb transceiver recently introduced by Northern Radio Company of Seattle, Washington. Rated at 5 watts peak envelope power, the MM-2C operates in the 1.6-10 MHz range. It's completely solid state and is powered by self-contained rechargeable batteries. The transceiver can be used with a variety of antennas and is protected against antenna mismatch. A temperature-compensated crystal oscillator assures optimum frequency stability between -22 and 122F (-30 and 50C).

The MM-2C transceiver is ideally suited for applications requiring portable communications capability in areas where the range with hand-held vhf fm equipment is limited by terrain or dense foliage. Although the unit is specifically designed for longer-range portable communications required by industries such as Petroleum, Mining, Geological, and Forestry, it is also suitable for amateur radio applications. For more information contact Northern Radio Company, 4027 21 Avenue West, Seattle, Washington 98199, or use *check-off* on page 126.

short circuit

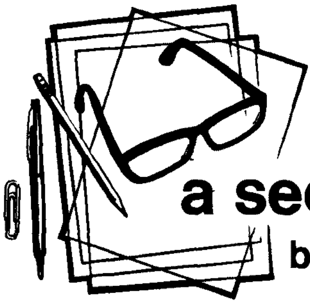
Much to our embarrassment, the so-called NASA speech filter which appeared in *circuits and techniques* in the June issue did not originate from NASA, but from a jokester who published the circuit in a French amateur radio magazine in April, 1971. The only clue to the spoof, which we didn't catch, was the name of the NASA engineer who supposedly developed the circuit — Schertz — which is German for joke. The only thing the circuit does is provide a very large insertion loss.

one dollar

ham radio magazine

OCTOBER 1976

second annual receiver issue



a second look

by Jim Fisk

In their quest for new sources of needed energy, researchers at the Lawrence Livermore Laboratory in California are putting together the world's largest laser in an effort to exploit the ultimate energy source – the same fusion energy as that used by the stars. The 200-foot (60m) long laser, which is scheduled to be completed by early 1977, is expected to produce 25-million watts of electricity. Initially, it will take more energy than this to make it work, so the system will operate at a net energy loss, but scientists expect to upgrade the mammoth to 100- to 300-trillion watts, and by 1980 it should make world history by generating more energy than it actually consumes. When that happens, it will signal the birth of the fusion age and the beginning of civilization's independence from scarce oil, gas, coal and atomic fuels.

Water, the fuel for fusion, has a fantastic amount of energy locked inside: the heavy hydrogen (an isotope called deuterium) in just 75 gallons of water could light New York City for nearly 10 minutes! Although scientists have known about this cheap and inexhaustible energy source for more than thirty years, harnessing it is no easy trick. Stars solve the problem by compressing the hydrogen and heating it to extreme temperatures, and then use their immense gravitational pull to confine the plasma so it can't escape. This works just fine if you're as large as a star (our sun is 864,000 miles in diameter), but here on earth the job will be done with a powerful laser which is carefully focused on a pinhead-sized pellet of heavy hydrogen fuel, instantly heating it to a tremendous temperature. Part of the pellet's outer shell is blasted outward; the inner part, however, implodes and compresses the fuel to 10,000 times its normal density. At the instant this happens the fuel fuses into lithium and emits a shower of heat and light. A thin layer of liquid lithium, which covers the spherical implosion chamber, carries away the heat which is used to drive turbines.

The huge laser, called *Shiva* after the Hindu god of destruction and reproduction, actually consists of twenty 1000-joule lasers, each about 150 feet (46m) long, arranged somewhat like a Gatling gun. Each of the individual lasers consists of a chain of seven elements: a master laser oscillator followed by six laser amplifiers which pumps out an intense pulse of light – equivalent to 10-million megawatts – lasting about a billionth of a second. Each of the 20 beams of carefully shaped and timed laser light simultaneously enter the ports of the implosion chamber and blast the fuel pellet into oblivion. The next pellet is then inserted into the center of the chamber and, once again, implodes after being blasted.

Although there are still a number of problems to be resolved, scientists are confident they will have the first successful fusion reactor in operation by early next year. It will take four or five more years to bring the system on line, but even at that, *Shiva* looks like the most promising new energy source to come down the pike in a good many years.

Jim Fisk, W1DTY
editor-in-chief

presstop

FIRST TWO METER WORKED ALL STATES was achieved by KØMQS when he worked K6YNB/KL7 via moonbounce August 17th. The Alaska contact was particularly timely for KØMQS as he'd just nailed his 49th state — Idaho — when he contacted W7UBI on August 2. The QSO with KØMQS was, coincidentally, K6YNB/KL7's first EME contact though the DXpedition station had made some Perseids contacts.

VHF/UHF Bands Continued to offer "DX" treats with long-haul contacts almost a daily happening on two meters during early August. W2BOC has asked for log data from any stations that participated in the June 26-28 openings on two and six meters to aid in sporadic E propagation studies he's making.

SALE OF GENERAL CLASS AMATEUR LICENSES is being checked by FBI in Indianapolis and possibly other cities. An article in the *Indianapolis Star* describes a supposed ring of Amateurs who were telling prospective Amateurs that "they can bypass rigorous testing procedures by purchasing a General Class ham license for $200" from the gang.

A "Mr. Big" in the operation is described in the article as "a government employee in Pennsylvania" who has access to the FCC's computers and, for a price, is "willing to feed the data into the computers to make sure a license will be issued."

The FBI's Agent In Charge in Indianapolis confirmed that an investigation is in progress but had no further details. Contacts at the FCC in Washington were not aware of either the story or the newspaper's report of it. Apparently a number of Indianapolis area Amateurs have been visited by the FBI investigators, and one of them said that an agent told him the Indianapolis investigation is "just the tip of the iceberg!"

FCC IS COMING DOWN HARD on bootleggers operating between the 27-MHz CB band and 10 meters. FCC agents and U.S. Marshalls nailed a dozen of the outlaw operators in a July raid in northern New Jersey, and seized an estimated $10,000 in radio gear.

Amateur Input On Unlicensed and/or out-of-band operators played a big part in the investigation, and FCC/Justice Department people in other parts of the country are also using reports from Amateur radio monitors.

Most Valuable Data is that providing general operating patterns of illegal operation in a given area. Recent discussions with key FCC enforcement people cited reports of specific frequencies, operating times and general location of operators as being of prime importance, with detailed dossiers on one or two violator's activities of lesser interest. Reporters should also specify what they use for monitoring and their location — information can go to any FCC Field Office or Monitoring Station.

"WN" NOVICE PREFIXES will no longer be issued after October 1, and Novice licensees will receive the same prefixes as higher class license holders. In its announcement of the change August 19th, the Commission cited processing problems as the prime reason for the change — its computer has had some difficulty in translating a WN into a WA, WB, or (recently) WD, resulting in the issuance of some duplicate calls.

Present Novice Licensees whose licenses expire after October 1 will also be changing prefixes — the Commission plans to issue them new licenses with the appropriate new prefix — until a new license is actually received, however, a presently-licensed Novice should continue using "WN."

WD Callsigns are being issued or are on the verge of being issued in the 2nd, 4th, 6th and 8th call areas, according to the Commission's release on the Novice change, and WDs are imminent in the 5th, 9th and Ø districts.

TWO LETTER CALL REQUESTS have totaled about 400 since July 1st with 75 or so of those from two-letter-call holders wishing to change. A few requests continue to trickle in, and the fourth district has very few 1x2 calls left. Though the next group of eligible Amateurs will use up some of those remaining, the next big influx of requests is expected in January. Don't be surprised to hear N prefixes and "X" calls such as W9XA by then.

PACKRATS' SOUTH AMERICAN MOONBOUNCE DXpedition in August was a resounding success with 16 stations in 8 countries worked off the moon on 432 MHz. First contact was with K2UYH, providing him with the first E-M-E WAC, followed by W3CCX, F9FT, I5MSH, PAØSSB, LX1DB, K3PGP, VE7BBG, W1JAA, JA1VDV, K8UQA, W4ZXI, W1SL, WØYZS, KØTLH and SM5LE. In addition to E-M-E operation, over 40 OSCAR 7 Mode B contacts were also logged before the group shut down to return to the States.

AMSAT'S ASCII STA has been granted and is good through February 6. More experimental use of ASCII through the satellites is being encouraged.

TOM McMULLEN, W1SL, has joined *Ham Radio* as managing editor. Tom, a well-known VHF/UHF experimenter, has been QST's Assistant Technical Editor.

Charlie Carroll, W1GQO, until recently a member of the ARRL technical staff, has also joined *Ham Radio* as an assistant editor. Welcome aboard!

optimum design for high-frequency communications receivers

High-performance circuits are described for communications receivers operating in the 10-kHz to 30-MHz range

The design of shortwave receivers has changed significantly in past years. Early receivers used double- or even triple frequency conversion, and selectivity was achieved at a fairly low frequency; for example, 455 kHz. Fig. 1 is a block diagram of this type receiver. Since the i-f had to be higher or lower than the frequency bands of these receivers, the i-f was usually set below 2 MHz, which resulted in image problems. To overcome this, many tuned circuits were required in the rf stages, and oscillator tracking became quite another problem.

Present-day design avoids these expensive mechanical arrangements thanks to the availability of crystal filters in the 30- to 120-MHz range. If an intermediate frequency higher than the highest frequency of reception is used, three advantages occur:

1. To achieve constant image and i-f suppression, a simple lowpass filter with a cutoff frequency of 31 MHz can be used, guaranteeing at least 80 dB suppression. Such a filter will also substantially reduce local-oscillator power radiation.
2. With both the oscillator and intermediate frequencies higher than the highest frequency of reception, gaps in receiver coverage are eliminated (a receiver with an i-f at 5.5 MHz, for example, precludes tuning between approximately 5 and 6 MHz).
3. The ratio of the oscillator frequency range (maximum/minimum) is, by definition, less than 2:1.

Additional input selectivity can be obtained by using fairly wide cross-modulation and intermodulation filters which operate in fixed-frequency bands. Thus the receiver can use bands of constant width; that is, in 1-MHz intervals.

If we talk about a good shortwave receiver, we must define "good" in terms of receiver electrical characteristics. Table 1 provides these characteristics for a shortwave receiver that is required for today's operation.

Many amateurs equate "good" with noise figure or signal-to-noise ratio. It has been recently agreed that a noise figure less than 10 dB is not essential for high-frequency amateur receivers. In some military and systems-oriented applications a noise figure of 6 dB or less is required for example, in systems using low-efficiency antennas such as whips. Unfortunately, sometimes some of this equipment is used with large antennas, so a large dynamic range in the receiver is needed despite the low noise figure.

What about oscillator radiation? The technique used by Southcom and being adapted by Atlas requires highly efficient shielding and filtering to avoid oscillator power feedthrough to the antenna input terminal. Commercially designed receivers for the military market must have a maximum of 15 microvolts reradiation of the oscillator signal — a requirement that practically none of the amateur receivers on today's market can fulfill.

new receiver design

Fig. 2 is a block diagram of a reciever with a first i-f of 40.525 MHz. This receiver covers 10 kHz to 30 MHz. The reason for choosing this i-f is that one of the standard i-fs in Europe is 525 kHz (455 kHz in the U.S.). By using a 40-MHz signal derived from an internal frequency standard, conversion from the first to the second i-f can be accomplished easily.

The signal at the antenna passes through a 30-MHz lowpass filter and, depending on the selection of the reception frequency, through either a high- or lowpass filter with a 2-MHz cutoff and one of eight automatically selected bandpass filters in the range of 2 to 30 MHz. The signal is then applied to a balanced power amplifier through an automatic attenuator circuit consisting of negative temperature coefficient and positive temperature coefficient resistors. The 30-MHz filter rejects image frequencies. The 2-MHz highpass filter separates the broadcast range with its often very high field strength from the high-frequency region and prevents BC signals from arriving at the mixer. An independent agc circuit (not a limiter) is used in addition to the normal agc system. This independent agc circuit

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

responds when signals equal to or above 100 mV are fed to the receiver input. This circuit acts as an automatic attenuator and protects all following stages from being overdriven without introducing any measurable distortion of its own. A push-pull power amplifier stage following the input filters uses heavy feedback to keep second- and third-order intermodulation distortion products as low as possible.

tween ±75 Hz to ±6 kHz, which permits optimum matching to the bandwidths required in the various modes of operation (slow CW, frequency-shift telegraphy or broadband telephony). Since individually switchable filters would hardly prove practical for such a large number of different bandwidths, a new double mixer with fixed-tuned filters has been designed (**fig. 3**).

The 525-kHz intermediate frequency with its side-

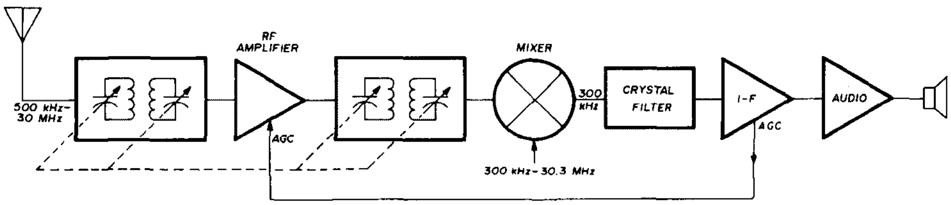


fig. 1. Block diagram of a single-conversion, low i-f, general coverage communications receiver typical of that used by many radio amateurs. The intermediate frequency is usually set below the tuning range of the receiver, as shown here, but this results in image problems on the higher frequencies.

The signal is converted to the first i-f in a high-level +17 dBm or higher double-balanced mixer and applied to a cross-modulation filter. This filter is a low-ripple, 6-pole crystal filter with a bandwidth of ±6 kHz or less, depending on its purpose. This filter suppresses adjacent signals, which might otherwise produce cross modulation in subsequent stages. A high-level double-balanced mixer is also used for converting the first i-f to the second *i*-f (525 kHz), where main selectivity occurs. The bandwidth of the second i-f stage is selectable in steps be-

table 1. Electrical characteristics for a modern shortwave receiver.

Frequency range	10 kHz to 30 MHz
Antenna connection	50 ohms, unbalanced
Preselection:	
10 kHz-2 MHz	lowpass filter
2 - 30 MHz	Bandpass filters for 2 - 3 MHz, 3 - 5 MHz, 5 - 7 MHz 7 - 10 MHz, 10 - 13 MHz, 13 - 17 MHz 17 - 22 MHz, 22 - 30 MHz
Setting accuracy	Better than 50 Hz; counter display 10 Hz, 1 Hz preferred
Noise figure	about 10 dB
Modes of operation	CW, MCW, a-m, dsb, ssb, afsk
I-f bandwidth	±75 Hz to ±6 Hz in steps
Shape factor	At least 1:1.45 (6/60 dB down)
I-f and image rejection	Greater than 80 dB
Crossmodulation	Less than 2% demodulation with 100 mV emf interfering signal and less than 10% with 5 V *emf interfering signal*
Blocking through second signal	3 dB with an unmodulated unwanted signal spaced 30 kHz from, and 100 dB above, desired signal of 1μV
Intermodulation distortion	
In band	>50 dB with two 50 mV emf tones
Second order	>90 dB with two 5 mV emf tones
Third order	>80 dB with two 5 mV emf tones
S meter	Should be calibrated in μV. Input voltage range <1 μV to >100 mV.
Outputs for accessories	Audio output 0 dBm Agc voltage for diversity reception I-f outputs Oscillator output for counter and digital programmer

band of ±6 kHz maximum is first converted into the range 52 to 64 kHz. It is then applied to a bandpass filter with a steep-edged selectivity characteristic at 64 kHz and converted to the original i-f. If the oscillator frequency is varied in the proper direction, the shift of the reduced i-f toward the steep filter edge clips or completely suppresses one sideband. Using the same oscillator frequency for mixing and remixing prevents any frequency errors from occurring between input and output, which might affect receiver setting accuracy. The position of the sidebands with respect to the carrier also remains unchanged. This arrangement is followed by a similarly designed selective circuit that can limit the other sideband. Similar selective filters are used in both circuits so that the oscillator frequency is symmetrical about the 525-kHz i-f. Thus, a frequency variation by the same amount but in the reverse direction results in a symmetrical bandwidth change. The selectivity at a 500-Hz spacing from the adjusted bandwidth is at least 60 dB, and the ripple within the i-f passband is about ±1.5 dB.

If filters with the same selectivity are used, double mixing ensures an identical, symmetrical skirt selectivity of the i-f passband characteristic, constant skirt selectivity for each bandwidth, and a phase delay characteristic symmetrical with the center frequency. With a-m reception, therefore, no distortion occurs due to delay time. With commonly used crystal or mechanical filters this high phase and gain linearity can't be achieved under any circumstances.

This method also provides a simple means of suppressing the upper or lower sideband during ssb reception. By setting the oscillator of the first and second selective circuits to a fixed frequency, variable limiting of the remaining sideband is possible.

Both conversion oscillators are synchronized with an internal reference oscillator to increase circuit stability. This avoids an impermissible variation or shift of the center frequency, especially in narrow bandwidths.

Advertised crystal filters, which are 8- to 10-pole

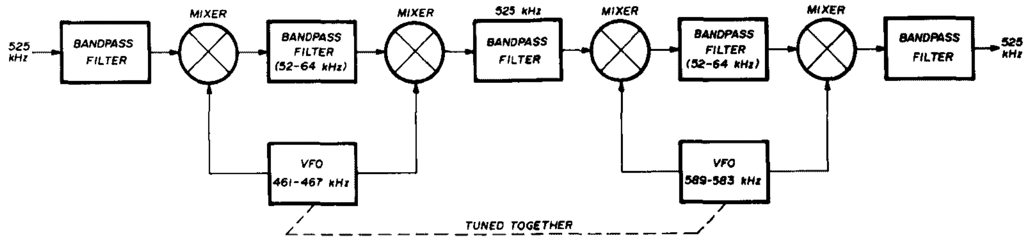


fig. 3. A system for obtaining variable i-f bandwidth through the use of a double mixing scheme. This system is used in the receiver of fig. 2 to provide selectable bandwidths from 6 kHz to 75 kHz without the expense of a large number of separate i-f bandpass filters.

designs with shape factors between 1.4:1 and 1.2:1, have two unpleasant effects:

1. The extremely steep skirt selectivity presents a problem for the agc circuit because of the high group delay and phase shift, which cannot be compensated. In almost all cases strong interfering signals at the edges of the filter response band will make the agc pump. This instability introduces distortion and overshoot.

2. Because of their high Q, and despite their fairly wide bandwidth, these filters produce appreciable ringing.

In addition, it's important to remember that, under the circumstances in which radio amateurs use their receivers, the i-f bandwidth for ssb reception should be between 1.9 and 2.4 kHz to avoid psychological fatigue caused by unpleasant hissing noises. More than ten years ago the bandwidth of the famous Collins KWM2 was restricted to 2.1 kHz for this reason.

I-f amplifier and gain control. The i-f amplifier boosts the incoming signal to an amplitude sufficient for distortion-free demodulation and ensures a constant output voltage by means of gain control. Its amplifying action is such that the inherent noise from the rf section is adequate to drive the succeeding agc amplifier to full output. If you analyze practically any amateur transceiver now on the market, you will discover that these receivers don't have enough i-f gain. The reason for this is obvious: most manufacturers want to avoid the expense of careful shielding, which is required if the gain is more than 80 dB. However, to obtain enough audio

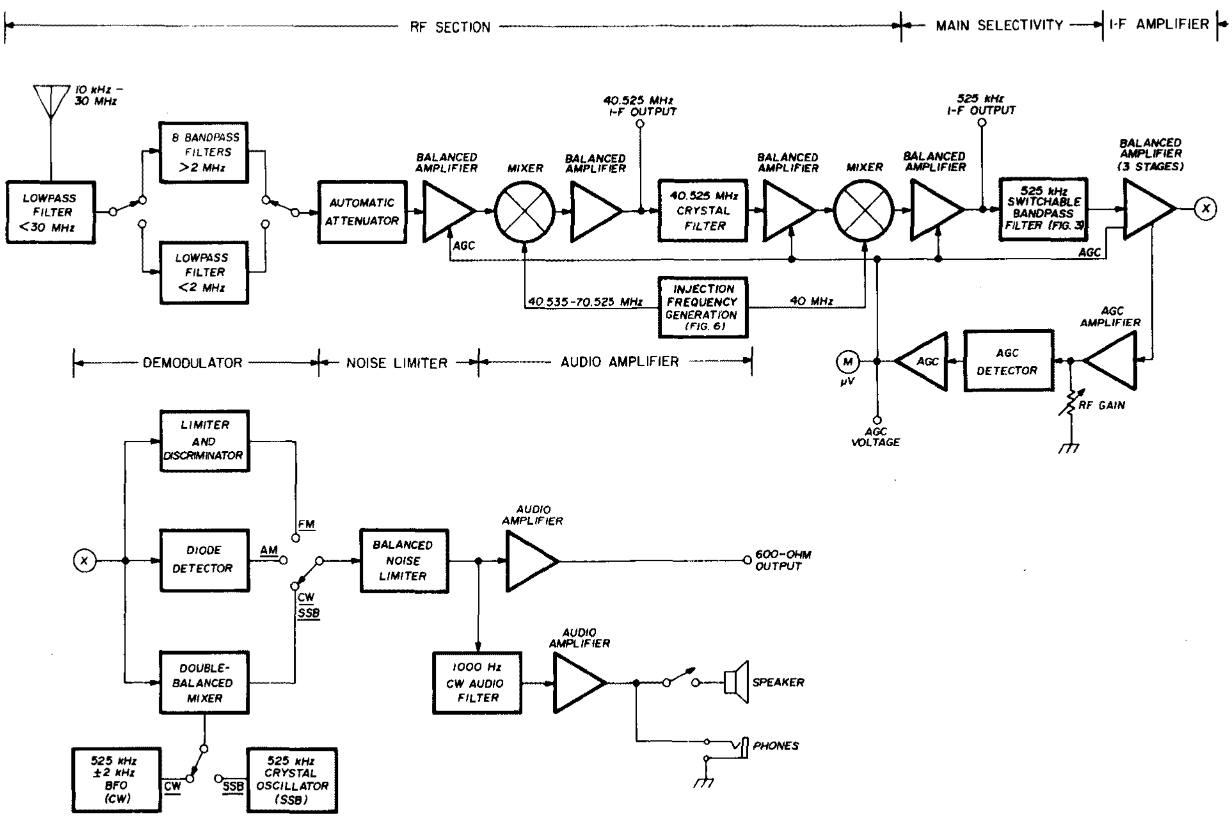


fig. 2. Block diagram of a high-performance, modern design communications receiver with an i-f at 40.525 MHz which overcomes many of the problems of previous designs. The noise figure of this receiver is about 10 dB, more than adequate for home-station use on any of the amateur high-frequency bands. Cross modulation, intermodulation distortion, and other operating characteristics of this receiver are listed in table 1.

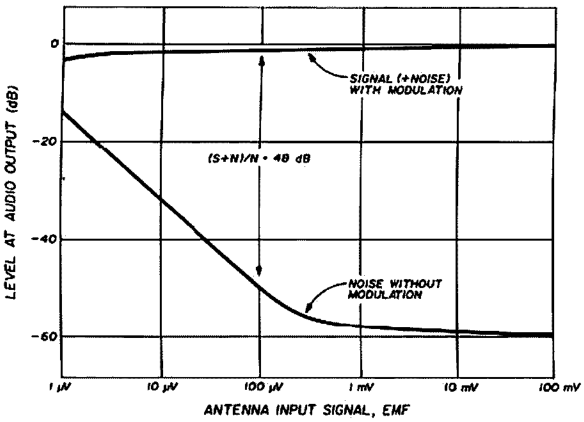
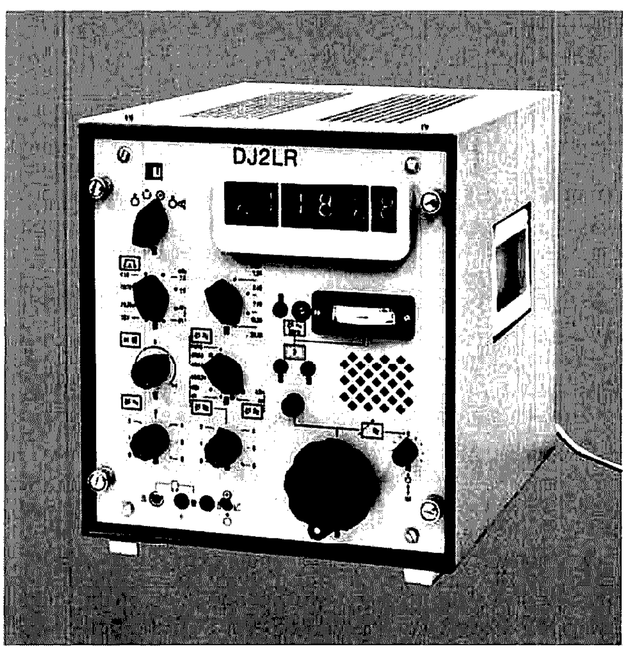


fig. 4. Optimum selection of agc action for perfect signal-to-noise performance.

output, a total gain of about 90 dB from the antenna input to the i-f output into the demodulator is required to obtain distortion-free demodulation. Unfortunately many designers use high-gain rf amplifiers and high-gain first mixers, both of which overdrive the second mixer. A sufficiently low overall noise figure can be achieved with little rf gain, as explained below.

Modern integrated circuits, such as the Plessey SL600 series or the Motorola MC1349 and MC1590, offer a good choice for a low-distortion, stable i-f amplifier with a minimum of shielding. Such a design permits sufficient agc to reduce unpleasant audio level changes. In addition, appropriate distribution of the agc voltage must be applied to the various stages to ensure a linear increase in the signal-to-noise ratio at small input voltages. Fig. 4 shows the agc response provided by the design discussed above. The agc circuit must have high gain; through pure control action input voltages up to 100 mV emf can be reduced to a residual error of ±1 dB without additional driving (forward-acting regulation). Furthermore, the agc voltage must be applied in appropriate levels to the various stages to ensure a linear increase in the signal-to-noise ratio at small input voltages. The result is a signal-to-noise characteristic much better than previously known.

Partially home-built receiver which tunes from 1.5 to 30 MHz. The PC boards were removed from a lab model Telefunken receiver, and the circuitry was redesigned using the techniques described in this article. The third-order intercept is +30 dBm, dynamic range is 120 dB or more, and the noise figure is 8 dB. Tuning accuracy is 1 part in 10 million.



The agc response time and the type of rectification must be different for each mode of operation. Peak rectification with a short response rise time of about 20 dB per 10 mS should be used for rapid leveling of the gain to the nominal value when receiving CW or ssb signals; a too-short response time does, however, lead to blocking by interference pulses. Average-reading rectification is recommended in the a-m mode to maintain receiver dynamic characteristics.

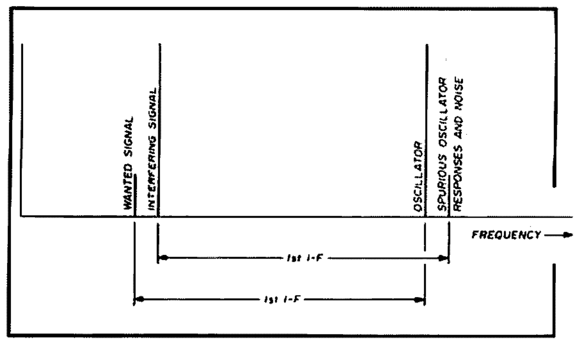


fig. 5. The noise sideband characteristics of the high-frequency injection oscillator affects receiver sensitivity because strong interfering signals at the mixer may convert the noise sidebands into the center of the i-f passband.

The agc voltage fall time should be selectable to obtain a gain variation between 5 and 50 dB per second. The slow response fall time maintains the dynamic characteristics and prevents any amplification of unwanted noise during CW pulse intervals. It also takes effect with a-m signals suffering selective carrier fading; it does not eliminate the distortion but prevents audio volume from increasing.

The modulation frequency of the incoming signal, which is present in the control lines during agc operation, must be filtered – without affecting the dynamic control characteristics – to such an extent that no additional distortion arises in a-m reception due to inverse modulation. In ssb reception, with intermodulation and little difference between the sideband frequencies ($f = 100$ *Hz*), intermodulation products ($2f_1 - f_2$) because of inverse modulation must be suppressed at least 50 dB.

Demodulator and af amplifier. The demodulators for the various operation modes are connected to the output of the i-f amplifier. A-m signals are rectified in a diode

fig. 6. The frequency synthesizer with linear master oscillator which is used in the high-performance communications receiver. A complete description of this circuit, which uses four separate oscillators, is contained in the text.

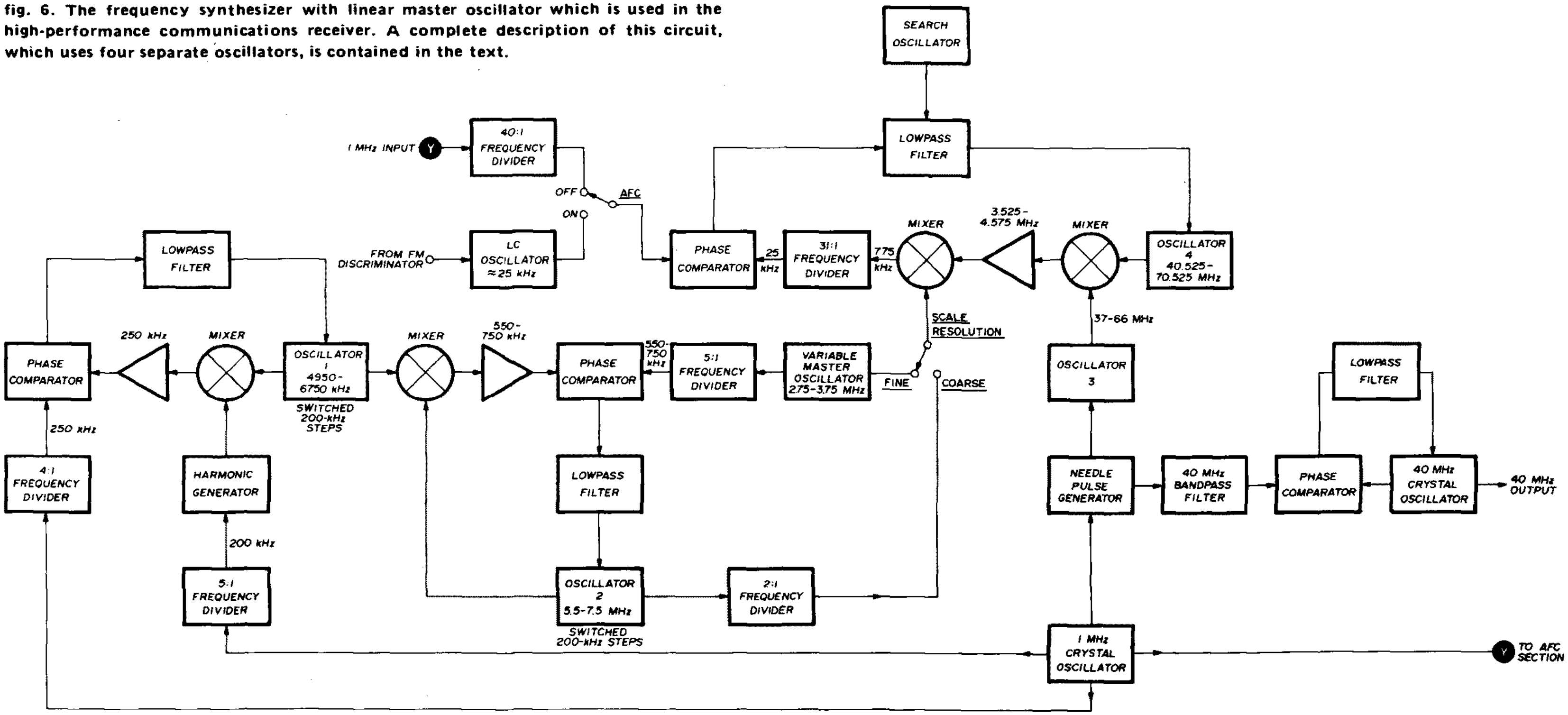
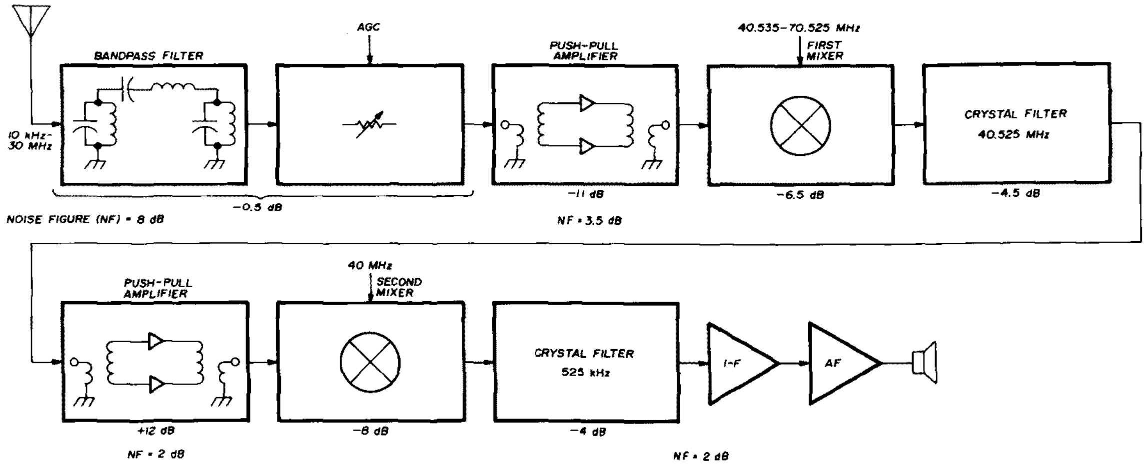


fig. 7. Gain distribution of a high-performance, high-frequency receiver. High dynamic range, as shown in fig. 8, is obtained by using as little gain as possible while keeping the overall noise figure below 10 dB.



circuit. The i-f, CW, and ssb signals are converted into audio signals in a double-balanced mixer which is driven by a vfo for CW signals, and by a high-precision crystal oscillator for ssb signals. Frequency-modulated waves (maximum bandwidth 6 kHz) are demodulated in a discriminator preceded by a limiter.

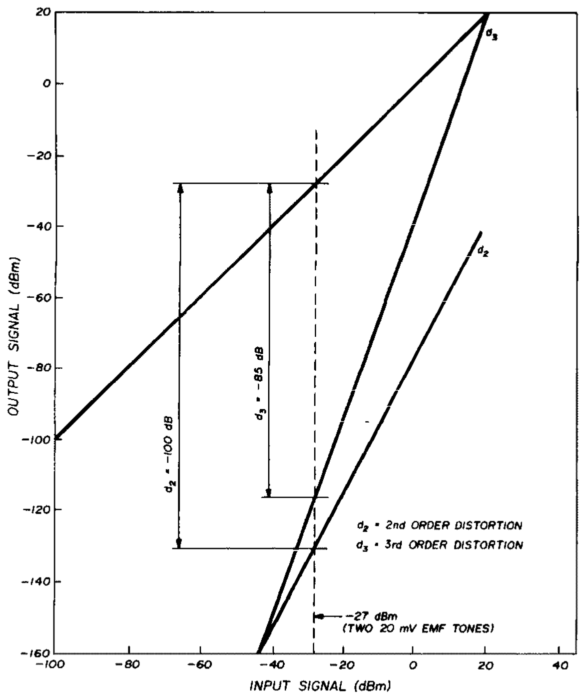


fig. 8. Dynamic performance of the receiver shown in fig. 7. The third-order intercept point is at +20 dBm. At −27 dBm input (two 20 mV emf signals), third-order products (d_3) are down −85 dB and second-order distortion (d_2) is down −100 dB.

The audio-frequency signal is applied through the function selector from the demodulators to an adjustable, balanced noise limiter, then to audio amplifiers. The amplifier with an output impedance of 600 ohms is associated with a balanced-line output. Starting from 0 dBm (1 mW into 600 ohms with 100% modulation), its output level is adjustable by ±10 dB. The second amplifier feeds a built-in speaker. It also contains a switch-selected 1000-Hz filter (bandwidth 200 Hz), which improves CW audio quality.

oscillator

The receiver frequency stability and setting accuracy depends on oscillator accuracy. The oscillator characteristics also affect receiver sensitivity, since strong interfering signals arriving at the mixer as a result of the receiver input broadband characteristic can convert noise sidebands and spurious oscillator responses into intermediate frequencies (fig. 5).

The oscillator frequency of the synthesizer (fig. 6), whose output is 40.525 to 70.525 MHz (oscillator 4), is

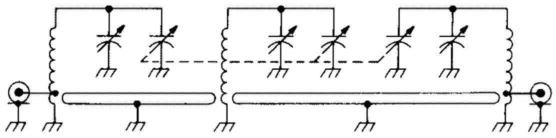


fig. 9. Tunable preselector which may be used to suppress local, high-level signals. Bandpass curve for this preselector is plotted in fig. 10.

derived in several phase-controlled circuits from selected harmonics of a 1-MHz crystal oscillator and the frequency of the variable-frequency master oscillator. The crystal in the 1-MHz oscillator is housed in a proportionally-controlled oven for stability. The frequencies derived from this 1-MHz oscillator are more accurate than those of the variable-frequency master oscillator, which determines receiver setting accuracy and stability.

While some receiver designs use frequency synthesizers which are switchable in steps down to 1 Hz, this much resolution is somewhat prohibitive for a search receiver. The lock time of such a frequency synthesizer would not permit easy tuning and, therefore, this type of receiver should only be used for channelized point-to-point communications. The method used here, derived from the Rohde & Schwarz EK07 shortwave receiver, was designed in 1957. A similar method has been used in receivers such as the National HRO500 and HRO600.

Basically, the 40.525 to 70.525 MHz frequency range is converted to an auxiliary i-f of 2.75 - 3.75 MHz. A frequency- and phase-sensitive double-balanced modulator is used as a comparator together with the master oscillator to determine oscillator 4 output frequency. The master oscillator, which can be tuned over a 1-MHz range (2.75 - 3.75 MHz) capacitor, exhibits a linear frequency characteristic. This linearity permits the use

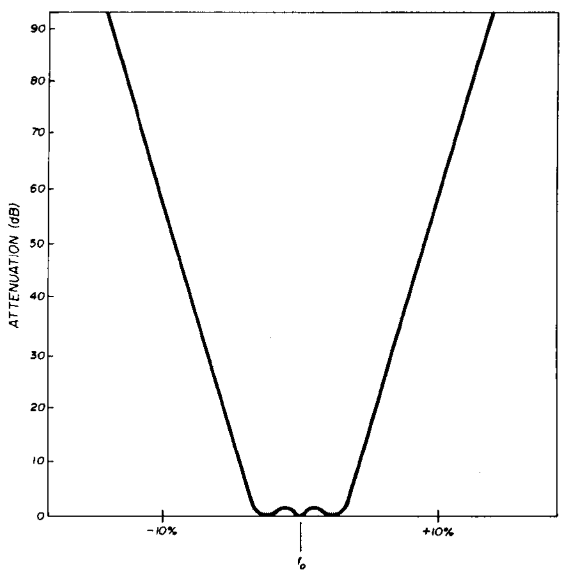


fig. 10. Selectivity curve for the preselector circuit of fig. 9.

of a directly calibrated scale with an ultimate resolution of 50 Hz.

To increase the variable master-oscillator frequency stability, all frequency-determining parts are housed in an oven and maintained at 60°C (140°F). Short needle pulses are derived from the 1-MHz oscillator output and are used to lock in oscillator 3 in 1-MHz steps. This frequency is required in a down-converter system to obtain 2.75 to 3.75 MHz for the master oscillator. The same needle pulses are used in a phase-locked loop system to synchronize the frequency of the 40-MHz crystal oscillator, which converts the first i-f (40.525 MHz) to the second i-f (525 kHz) with the 40th harmonic of the 1-MHz crystal oscillator. Oscillator 4 consists of six single oscillators, each of which is tuned over a range of about 5 MHz by tuning diodes.

The frequencies of oscillators 3 and 4 are converted to 775 kHz. This frequency is then applied to a 31:1 frequency divider. A phase comparator operating on the scanning principle compares the 25-kHz frequency difference with the output from the 1-MHz crystal oscillator. This output voltage adjusts oscillator 4 (residual superimposed ac voltages have been suppressed by the lowpass filter). If this circuit is not controlled, as for instance when the MHz range is changed, a search oscillator generates a sawtooth voltage that sweeps the oscillator range until its frequency is in the lock-in range of the control circuit. The frequency divider reduces the control circuit gain to the 31st part, thus increasing stability.

When the afc circuit is switched on, the 25-kHz reference frequency is not derived from the 1-MHz crystal oscillator but from an LC oscillator, whose frequency can be varied by a varactor. If the receiver is not

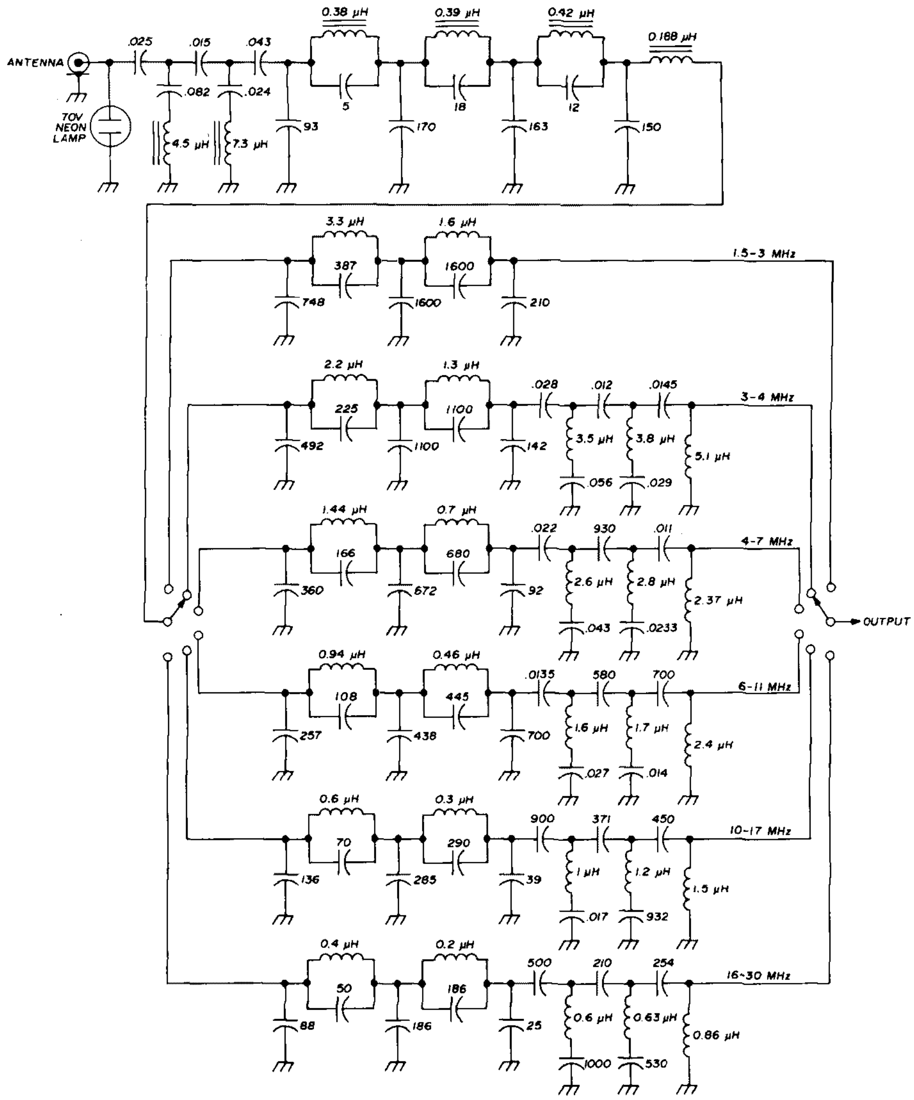


fig. 11. Highpass-lowpass elliptic filter (1.45 MHz to 32 MHz) for use at the front end of a high-frequency communications receiver.

exactly tuned to the carrier, the output voltage of a narrowband fm discriminator at the output of the i-f section changes the LC oscillator frequency and hence that of oscillator 4, so that the carrier lies at midband except for a residual control error of about 5% of the original deviation. The afc circuit permits adjustment to within ±1.5 kHz.

Free-running oscillator 4 has a high signal-to-noise ratio; for example, more than 140 dB per hertz of bandwidth at a 30-kHz spacing from the carrier. The control circuit has a narrow bandwidth to minimize its effect on the signal-to-noise ratio (circuit gain of 1 at about 5 kHz). Low-frequency disturbances, such as shock or noise caused by operating the switches or by the built-in monitoring loudspeaker, are avoided by mounting parts of the oscillator on shockmounts.

The mode of operation described is ideal for rapidly scanning wide frequency ranges, since a range of 1 MHz is covered in ten rotations of the tuning knob. The mode with high-scale resolution provides the full setting accuracy of the receiver. For this purpose, the tuning knob for the 100-kHz ranges is turned from its end position to the required 100-kHz step. The signal fed to the control circuit of oscillator 4 is a mixture of the tenth part of the master-oscillator frequency and a frequency derived from the 1-MHz oscillator. This signal is switch-selected in 100-kHz steps.

A frequency divider and a harmonic generator produce a 200-kHz spectrum from the 1-MHz frequency of the oscillator. Oscillator 1, like oscillator 2, is switched in 200-kHz steps by the 100-kHz tuning knob and converts one of the spectral lines in the mixer to 250 kHz. The subsequent phase bridge compares this frequency with a 250-kHz frequency derived from the 1-MHz oscillator by a 4:1 frequency divider, thereby producing the control voltage for oscillator 1. Oscillator

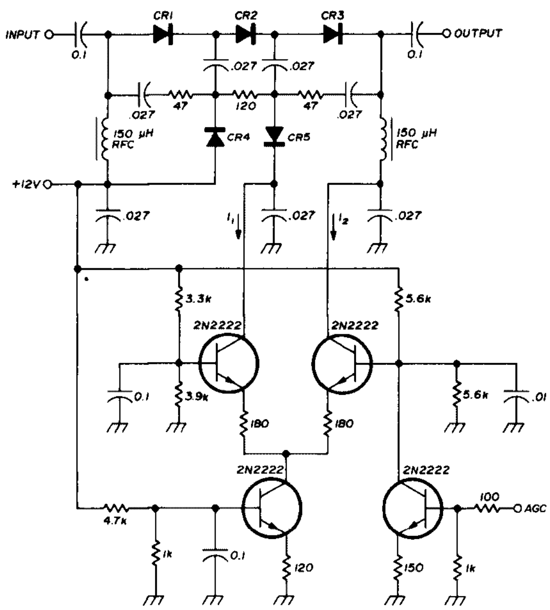


fig. 12. PIN diode attenuator with dc amplifier for use in a high-performance communications receiver. All diodes are Hewlett-Packard PIN diodes, type HP5082-3081.

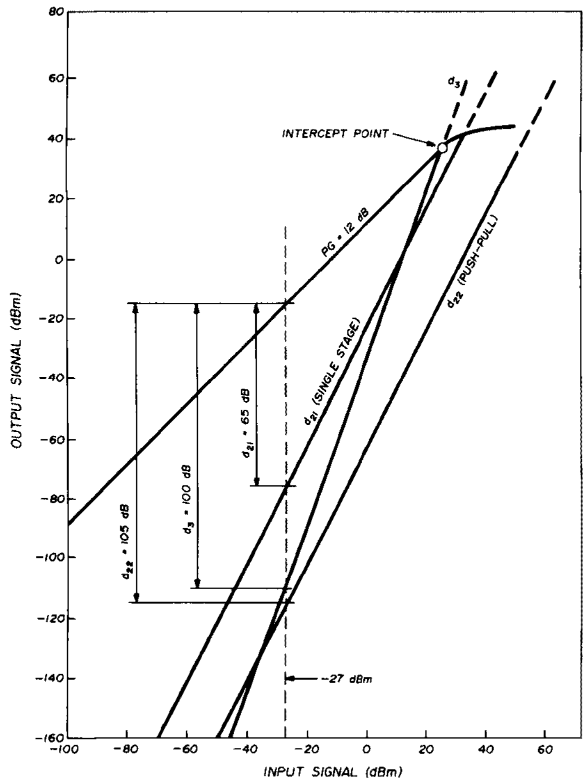
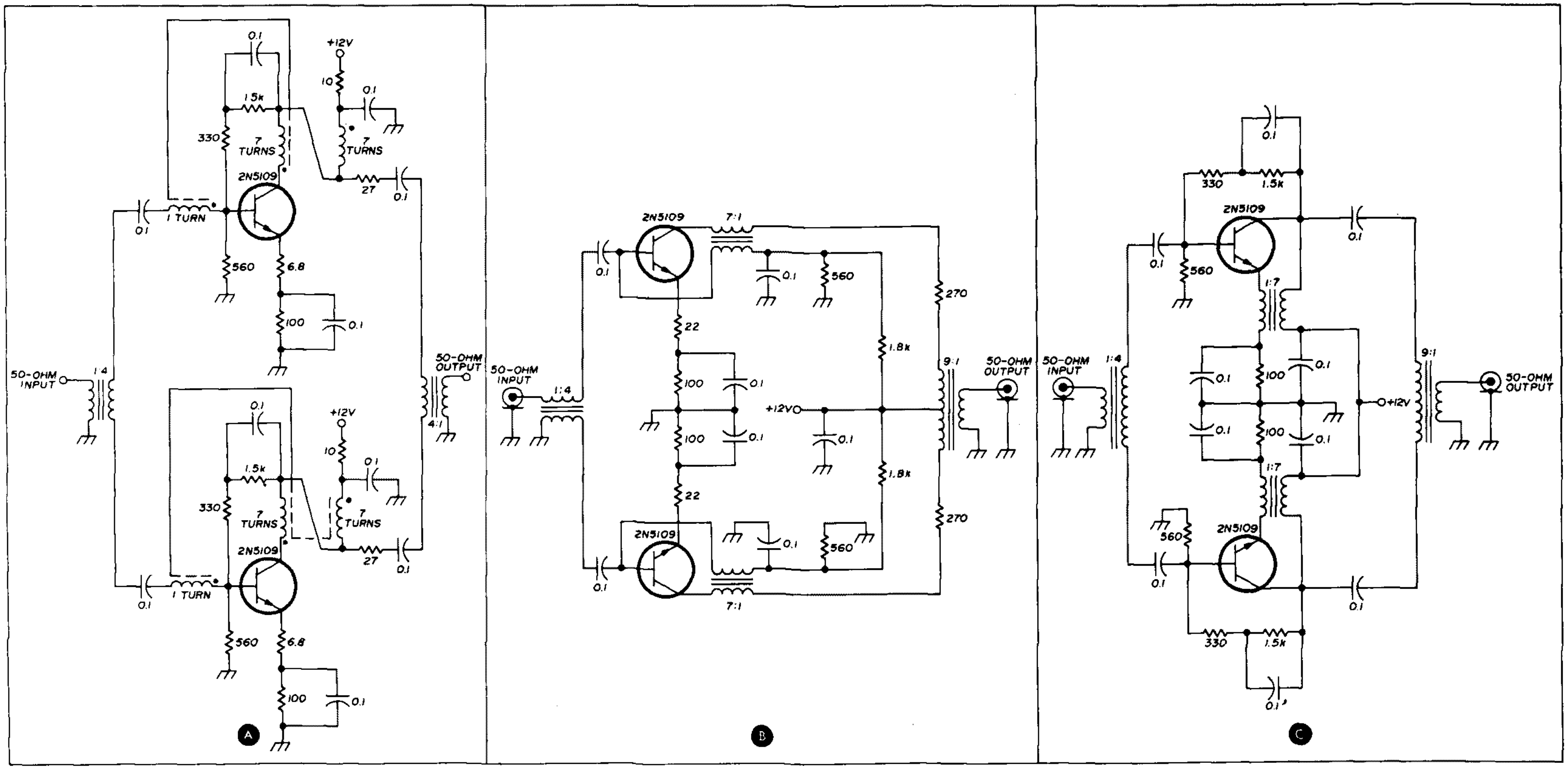


fig. 13. Dynamic performance of a new type of push-pull rf amplifier stage (figs. 14A, 14B, and 14C). Power gain is about 12 dB. With an input of −27 dBm (two signals), third-order distortion products are down 100 dB and second-order distortion is down 105 dB. Third-order intercept point occurs at an input of about +32 dBm

1 thus operates with the high accuracy of the 1-MHz crystal oscillator on one of the frequencies divided into 200-kHz steps between 4950 and 6750 kHz.

In a further phase-control circuit, the frequencies of oscillators 1 and 2 are converted to the 500 - 700 kHz range and compared in the phase bridge with the master oscillator frequency, which is divided in a ratio of 5:1. After separating the ac voltage components in the low-pass filter, the output voltage of the phase bridge controls the frequency of oscillator 2. By dividing the oscillator frequency in half, the frequency of the master oscillator, 2750 - 3750 kHz, is again obtained, but this time in a switch-selected, 100-kHz range. Because of the frequency divisions (5:1 and 2:1), the master oscillator covers only a tenth of the range so that receiver scale resolution and accuracy are increased by ten times. With a counter, the scale resolution may be 10 Hz or better and the BCD outputs may be used to program a direction finder through a remote-control system.

This type of synthesizer permits continuous tuning for many purposes. However, quasi-continuous tuning in steps of 10 Hz is admissible, which simplifies the whole synthesizer dramatically. This synthesizer has been described in detail to show what must be considered when building a practically spurious-free, low-sideband-noise local oscillator.



A. Circuit using current and voltage feedback for improved linearity and transformers for stabilizing input and output impedance from 100 kHz to 200 MHz. Amplifier has low output impedance; the two 27-ohm resistors are used to increase this value.

B. An improved version of the circuit in A. Output impedance is very high. Agc can be applied by replacing the two 270-ohm resistors with a single pin diode shunt regulator. This circuit is less expensive and simpler than a constant-impedance, T-attenuator.

C. Low-noise version with emitter feedback for extremely high input and output impedance. Amplification is 7 (1:7 turns ratio) and input impedance is 300 ohms divided by 7, or 47 ohms. Output impedance is several hundred ohms. Noise figure of less than 2 dB can be obtained.

fig. 14. Three push-pull arrangements for high linearity input rf stages.

The receiver described above is the result of combining various optimized stages and careful planning. To predetermine the overall technical performance, it's required to start with a block diagram containing vital information, such as gain, dynamic range expressed in second, third, and higher-order intermodulation distortion, and noise figures.

Fig. 7 is a block diagram of a receiver with a first high i-f and shows the amplification or losses of each stage. As shown, the input bandpass filter and the automatic attenuator have a 0.5-dB loss. The push-pull rf amplifier provides approximately 11 dB gain, which will compensate for the losses of the passive, high-level, double-balanced mixer and the losses of the 40.525-kHz crystal filter. The overall gain from the rf input to the 40.525-MHz push-pull amplifier is 0 dB. The stages from 40.525 to 525 kHz are almost identical; however, the second-order intermodulation of the second rf amplifier does not exceed the high values of the first. This is not necessary because of the cross-modulation crystal filter. The second high-level double-balanced mixer has somewhat higher losses but provides the same or even better intermodulation distortion suppression. The first 525-kHz filter is about ±6 kHz wide, and the overall gain up to this point is 0 dB.

The overall noise figure, which can be calculated from the block diagram, is 8 dB. The second and third intermodulation distortion performance of the rf input is shown in fig. 8. The main idea of obtaining a high dynamic range lies in the concept of having as little gain as possible while keeping the overall noise figure below 10 dB. This is achieved by carefully selecting the characteristics of each of the individual stages, as will be described later.

For a few rare applications it may be necessary to suppress frequencies in the immediate vicinity of a desired signal, for example, where a transmitter and a receiver are used simultaneously. Under these rare circumstances an rf preselector, as shown in fig. 9, is an absolute necessity. Because of the preselector's high selectivity, a 10-volt emf signal, ±10% away from the frequency of reception should not create intermodulation distortion more than -9 dB down. However, for these purposes special transceivers with duplex capabilities are used. These preselective filters have about -4 to -5 dB insertion loss. A typical response is shown in fig. 10.

circuit analysis

Because of the new approach in the design of this receiver, the circuits of a number of stages will differ substantially from those commonly used. To give a better understanding of the overall system, the most important stages are shown and explained here.

Input filter. In most cases the highly selective preselector of fig. 9 is not required, so a combination of elliptical filters selected for the frequency ranges is adequate. Third- and higher-order, odd-number intermodulation distortion products cannot be overcome by selectivity unless filters are 10 kHz wide or less. Second-order

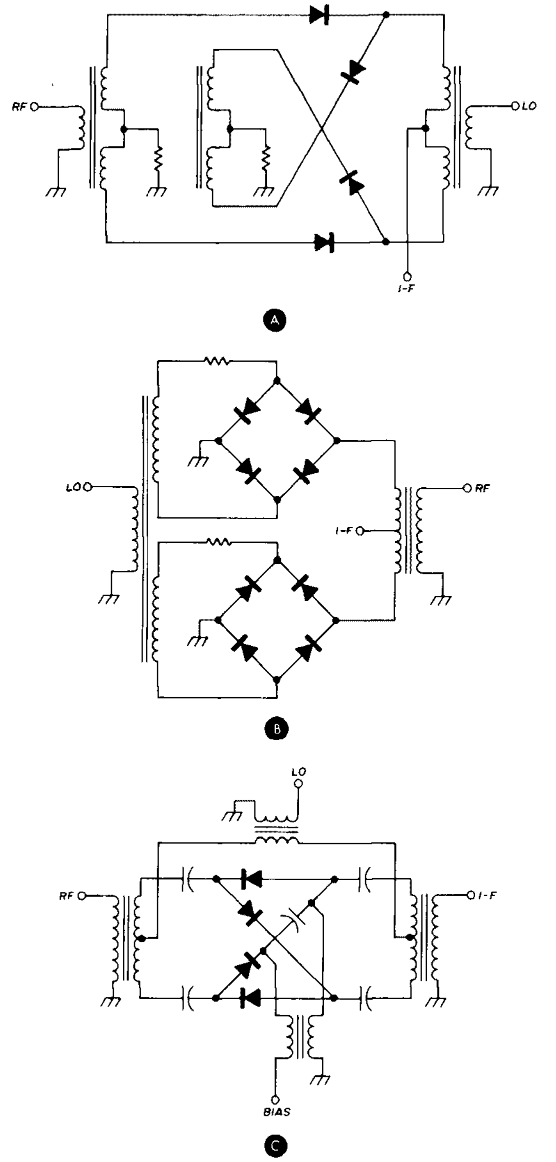


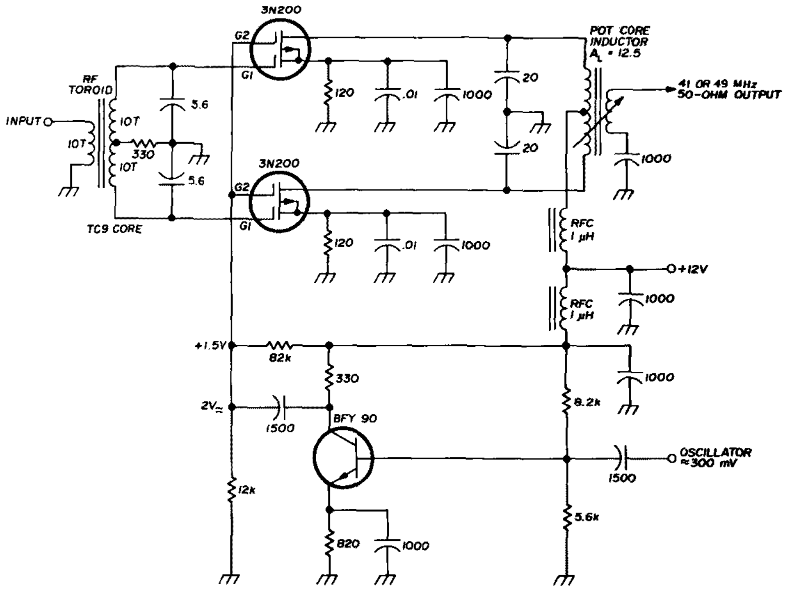
fig. 15. Double-balanced mixers. A and B are medium- and high-power configurations; C is a double-balanced mixer for uhf/shf with adjustable bias.

intermodulation distortion products can be improved by about 40 dB with these switchable filters.

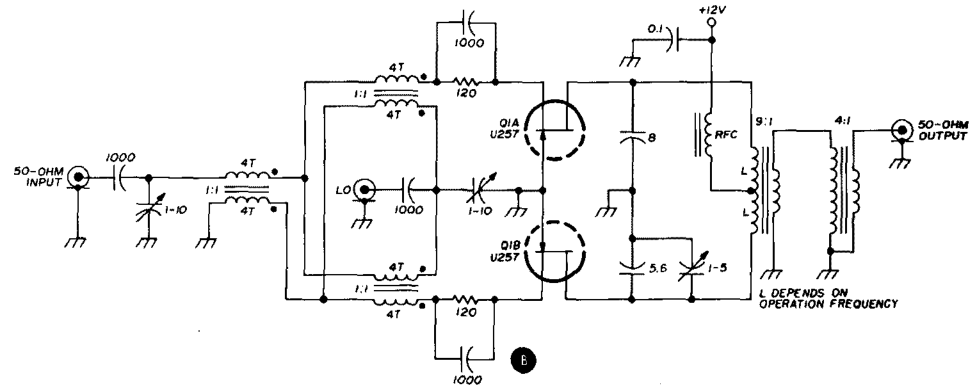
Special Bessel-Cauer elliptical filters are used with a given Chebishev response in the passband. This is absolutely necessary, because 50-ohm impedance matching is required in the stopband and passband so that filters can be cascaded. Fig. 11 shows a modern input filter covering the range 1.5 to 30 MHz in six steps.

A highpass-lowpass input bandpass filter suppresses unwanted broadcast signals, and the lowpass filter section, together with the following lowpass section of the individual bandpass filter, guarantees more than 90 dB image suppression. The advantage of this technique is that these filters can be aligned very easily with the help

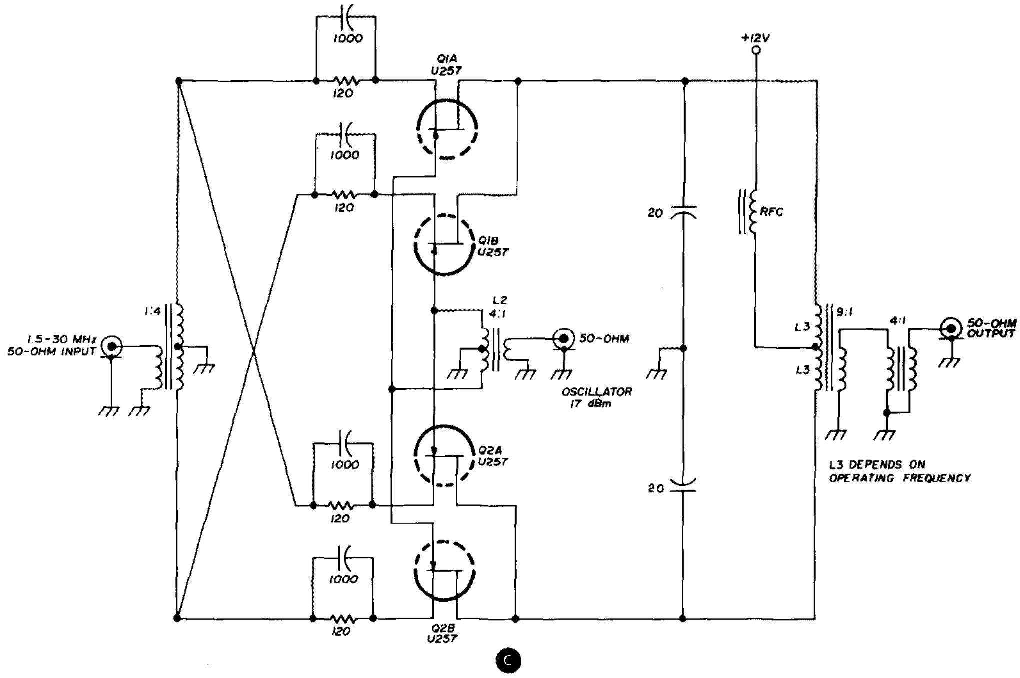
fig. 16. Examples of fet mixers. A balanced mixer with high input impedance (1 kilohm) is shown in A. A two-tone, 176 mV emf signal produces third-order IMD 68 dB down. A balanced fet mixer with 50-ohm input is shown in B (same performance as the circuit in A). C is a double-balanced mixer suggested by Ed Oxner of Siliconix. Its performance is 3 dB better than that of the circuit in B. An improved version of this mixer is shown in D (on facing page), also by Mr. Oxner.



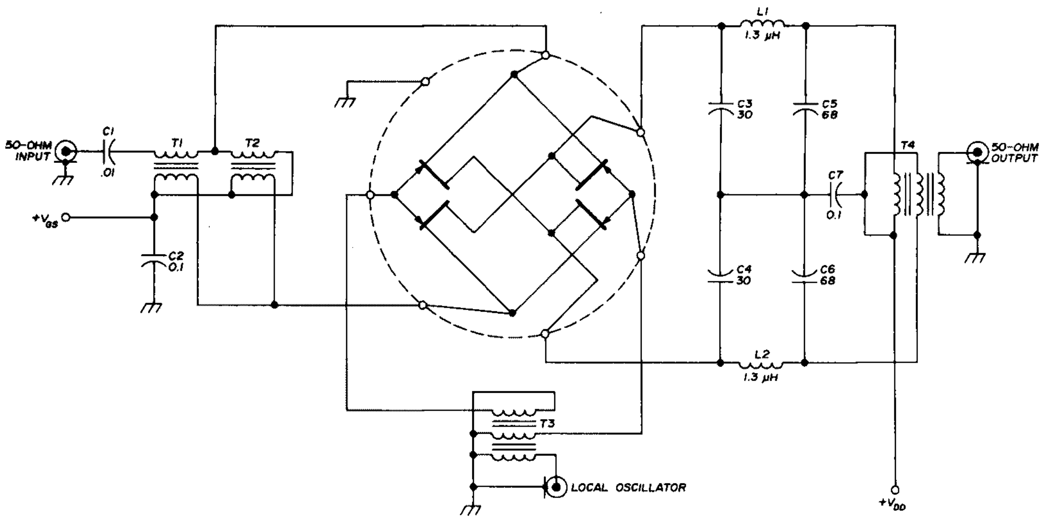
A



B



C



D

of a sweeper generator and an oscilloscope, and are much less complex than a mechanical tracking arrangement.

Input attenuator. Modern communications receivers should have low-distortion, automatic input attenuators that are activated at input signal levels above 100 μV emf. For the frequency range from 1 MHz to 30 MHz, pin diodes such as the HP5082-3081 are highly recommended in a double-T configuration (fig. 12), where the line impedance must remain fairly constant because of filter matching. The intermodulation distortion products of an attenuator such as this are about 85 dB down for two 1-volt emf signals. This means that the attenuator does not add any serious distortion to the system.

Rf input stage. In the past only certain vacuum tubes were considered to be sufficiently linear for rf front ends. However, recently it has been found out that if both voltage and current feedback are used in a transistorized rf stage, much better linearity can be obtained than with any receiving-type tube. The second-order intermodulation distortion products can be suppressed by almost another 40 dB over a single stage when using a push-pull arrangement. Fig. 13 shows the intermodulation distortion performance of a single stage and a push-pull stage, using the highly linear 2N5109 uhf power transistors made by RCA. The input and output impedances are stabilized with a third feedback network using a wideband transformer.

Fig. 14 shows three push-pull arrangements suitable for this purpose. Fig. 14A, which was published earlier,[2,3] uses voltage and current feedback to minimize intermodulation distortion, while the transformers act as impedance stabilizing devices. The cores recommended are F625-0-TC9 for 1.5-30 MHz or F625-9-Q1 for the higher frequencies (made by Indiana General).

Fig. 14B shows an improved version of the circuit in fig. 14A. While the latter produces a 50-ohm output impedance, the circuits of figs. 14B and 14C are basically constant-current devices. Because of this, a push-pull pin diode attenuator using only two diodes can be used (5082-3081 made by Hewlett-Packard). With the constant-impedance attenuator, fig. 12, second-order intermodulation distortion products are also further suppressed. However, because of the unbypassed emitter resistors, the additional noise contribution will not permit noise figures below 3 to 4 dB.

Fig. 14C show a low-noise arrangement in which

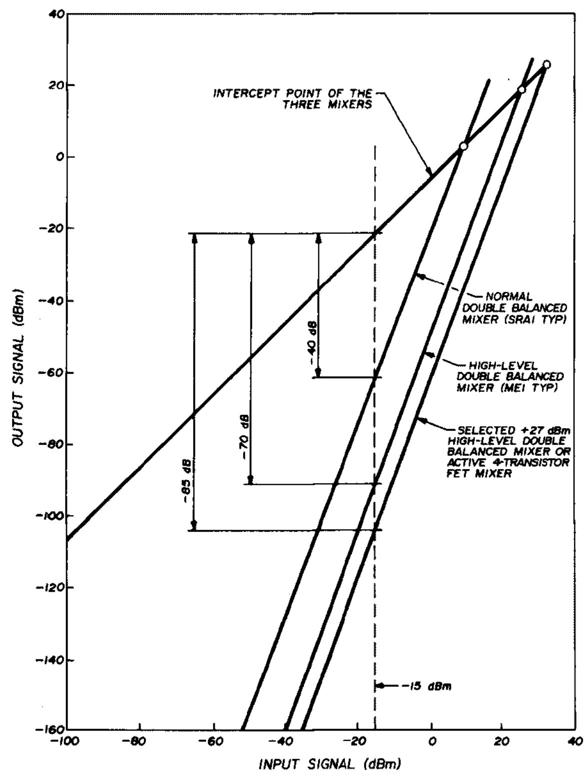


fig. 17. Comparison of three hot-carrier, double-balanced mixers showing third-order IMD performance.

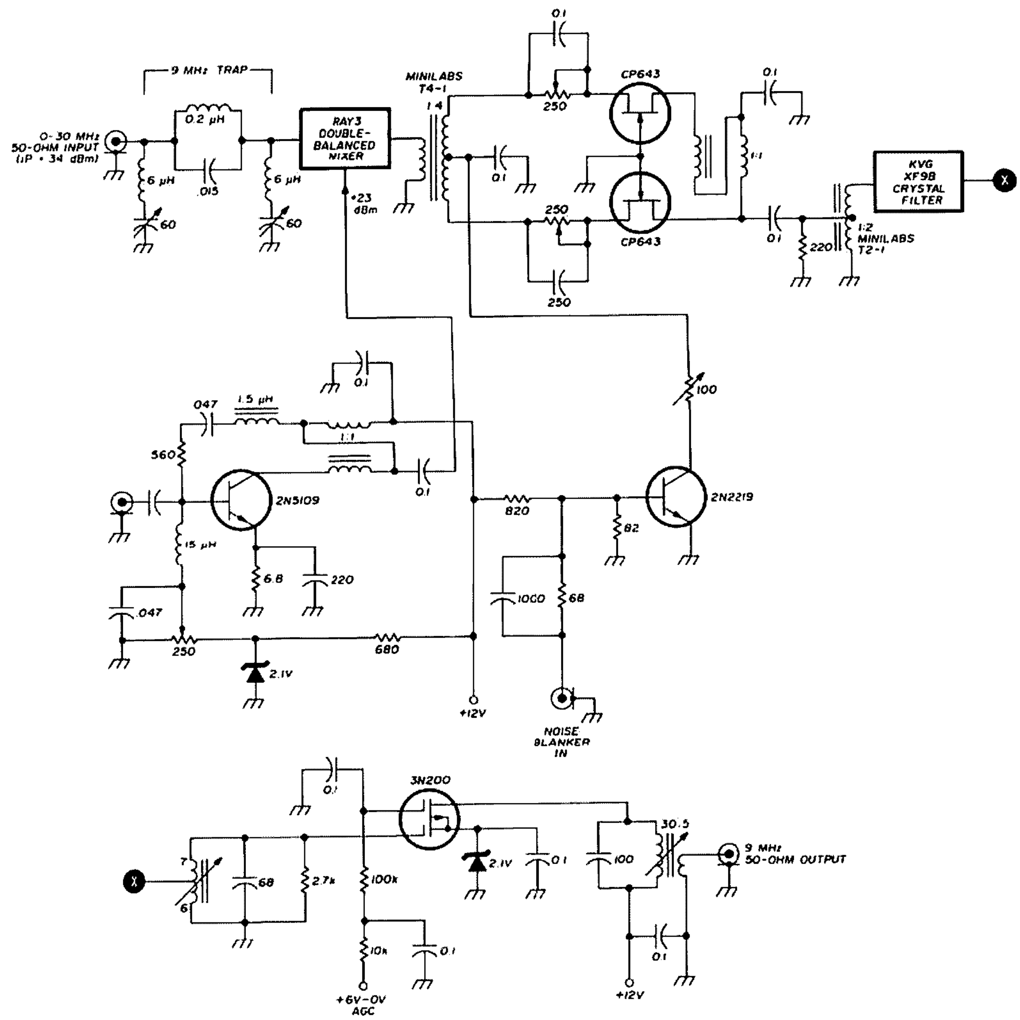
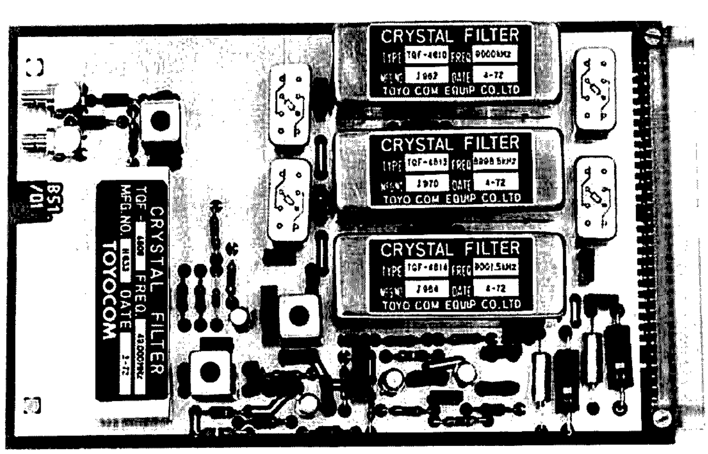


fig. 18. Improved version of Martin's rf input stage. The 9-MHz trap suppresses i-f feedthrough, and the push-pull fets increase dynamic range by 4 dB relative to the original design. In addition, a noise blanker can be used with 1-μs switching time because of the push-pull arrangement.



I-f section of the high frequency receiver built by DJ2LR. Crystal filters are switched into the circuit with small relays. The i-f amplifiers are located along the bottom portion of the board.

emitter feedback through a transformer is accomplished. As a result, the input and output impedances are extremely high, and the voltage feedback permits choosing suitable values. However, this push-pull stage will provide noise figures of 2 dB or less, while still having the same basic dynamic range.

In some cases it's possible to build a receiver without an rf input stage. Especially for fixed operation, the antenna system will have a noise temperature much higher than 300 KT_0 , which means that the various noise sources being picked up by the antenna system will provide a noise floor significantly higher than the receiver-input noise. Under these circumstances, receiver sensitivities up to 20 dB do not degrade overall performance. For applications like mobile radios, however, such sensitivity is not good enough. Since mobile antenna efficiency is between 10 and 30%, the noise pickup is reduced by the same amount and, therefore, noise figures less than 10 dB are vital. In cases of fixed stations new circuitry can be used, which is discussed in the next section.

Double-balanced wideband mixers. It has been found that push-pull arrangements have advantages over single-stage mixers; however, because of the high input impedance of tubes, few attempts have been made to use push-pull or double-balanced mixers with vacuum tubes in high-frequency communications receivers. Because of some other disadvantages the use of beam-deflection tubes was not a breakthrough. Since the introduction of low noise hot-carrier diodes, double-balanced mixers for high-level operation have been constructed and various configurations have been used, as shown in fig. 15.

Only recently double-balanced mixers with field-effect transistors have been used which exhibit excellent third-order intermodulation distortion suppression. Fig. 16 shows four tested configurations with fets. Figs. 16B and 16C use matched pairs.

Fig. 17 shows the third-order intermodulation distortion suppression of low-level double-balanced diode mixers, medium-level double-balanced mixers, and high-level double-balanced mixers, including double-balanced, field-effect transistor mixers. As can be seen, the fet mixer shows significantly higher signal-handling capability. Most circuits using field-effect transistors in double-balanced mixers were originated by Ed Oxner (ex W9PRZ) of Siliconix Incorporated.

Because of the inherent gain capability, fet double-balanced mixers offer advantages in test instruments but their application in shortwave receivers remains debatable. These circuits are very sensitive to load termination, and the input impedance of most crystal filters found on the market changes significantly over frequencies outside the passband. To reduce this effect, a resistive termination is required at the input (valid for all mixers, independent of configuration). However, the remaining impedance jump of 1 to 2 for out-of-passband operation reduces performance greatly.

Field-effect transistors are high-input-impedance devices. Fet mixers require about 2 volts rms across 50 ohms, which is equivalent to the injection required by a high-level double-balanced mixer. Because of the difficulty in providing wideband impedance matching in fets at high impedances, in my opinion medium- to high-level double-balanced mixers offer an advantage over fet mixers.

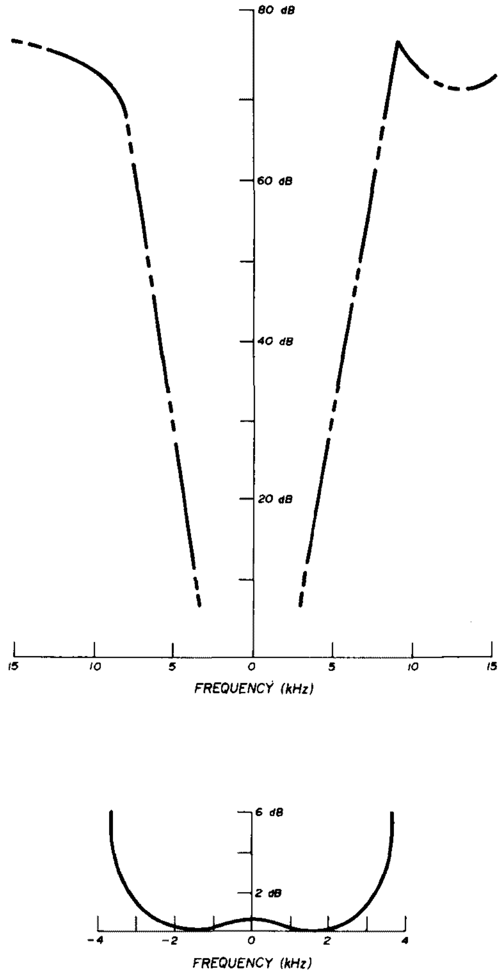


fig. 19. Selectivity curve of a 41-MHz crystal filter manufactured by Toyocom in Japan (Toyocom TQF4633, about $80).

A typical application for an input stage using a double-balanced mixer and no rf stage is shown in fig. 18. This is an improved version of a circuit designed by Mr. Martin, based upon an article I published in 1972.[4] The 9-MHz trap in the input suppresses image feed-through. The grounded-gate operation of the CP643 transistors (Crystalonics) provides a wideband resistive input termination for the mixer. The magnitude of the input impedance to the CP643s can be set by adjusting the 250-ohm potentiometers.

The 2N5109 transistor amplifier provides about 20 dBm injection for the RAY3 high-level double-balanced mixer (Mini-Circuits Laboratory). This circuit, which is

similar to that found in the Atlas transceiver, has a distinct advantage: because of proper termination of the double-balanced mixer, unwanted intermodulation distortion products are at least 15 dB below those found in the Atlas circuit. The reason for this is that the Atlas circuit uses a tuned circuit between the crystal filter and the first i-f transistor (2N3866), and this tuned circuit does not provide proper matching to attenuate spurious products. All manufacturers of double-balanced mixers

this push-pull amplifier is 220 ohms and a crystal filter requires 500 ohms, a 2:1 transformer, model T2-1 (made by Mini-Circuits Lab), must be used.

The 220-ohm resistor reduces the high-impedance characteristic effect of the filter as described in reference 3. The output termination of the crystal filter is provided by the tuned circuit, which is heavily damped by the 2.7 kilohm resistor. This is necessary to provide stable operation.

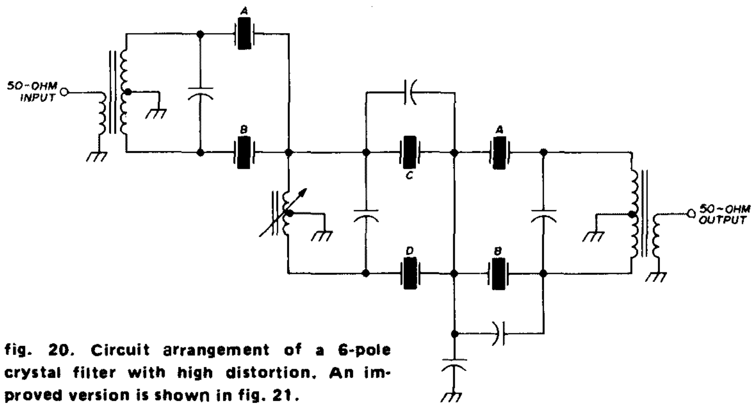


fig. 20. Circuit arrangement of a 6-pole crystal filter with high distortion. An improved version is shown in fig. 21.

specifically require proper resistive termination for optimum performance.

The KVG XF9B crystal filter is a popular ssb filter familiar to most amateurs. The first i-f stage after the crystal filter shown here provides enough gain and agc action for most applications.

The RAY3 double-balanced mixer requires a 50-ohm termination, which is obtained by an rf transformer, model T4-1.* Using a push-pull arrangement not only increases total dynamic performance by 3 dB but also provides the possibility of using the transistors as a switch for noise blanking. For high-efficiency noise blanking a switch time of 10 microseconds is essential. A

The field-effect transistor provides 60 dB gain variation, and the overall gain from input to output at 50 ohms is 28 dB. In cases where a tuned circuit is used as a first i-f amplifier between the double-balanced mixer and an input stage, the dynamic range of the mixer will be heavily degraded. This is one of the reasons why the Atlas circuit does not provide the dynamic range that would be expected from the mixer alone. I believe this circuit is a good suggestion for experimenters who are still willing to build their own communications receivers.

Vhf crystal filters. Until recently low-loss 6-pole crystal filters were not available. The most significant feature

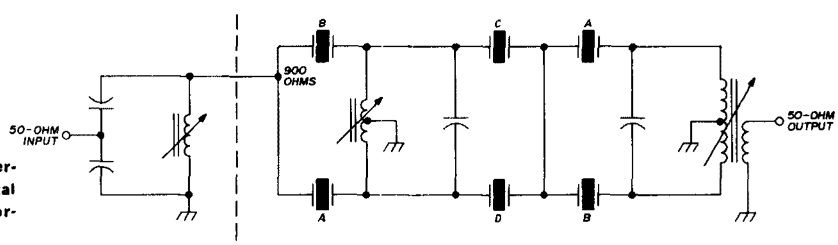


fig. 21. improved version of a 6-pole crystal filter with low distortion.

similar circuit, which is extremely efficient, is found in the Swan 200 series of amateur equipment.

To reduce costs, or if the CP643 is difficult to obtain, two BF246C transistors, made by Texas Instruments, may be substituted. These are high-current vhf fets with an I_{DSS} of about 100 mA. Since the optimum load for

*The RAY3 mixer and T4-1 transformer are made by Mini-Circuits Laboratory, 837-843 Utica Avenue, Brooklyn, New York 11203.

other than the shape factor of these vhf filters is the internal loss, which must be held below 5 dB (typical 4.5 dB) for narrow-band applications, i.e., ±3.5 kHz. Fig. 19 shows the typical response of a 41- or 49-MHz crystal filter.

When it was discovered that mechanical filters with magnetostrictive transducers create heavy intermodulation distortion at high input voltages, these filters were updated and now use piezo-electric transducers, which avoid this distortion. A similar effect can be observed at

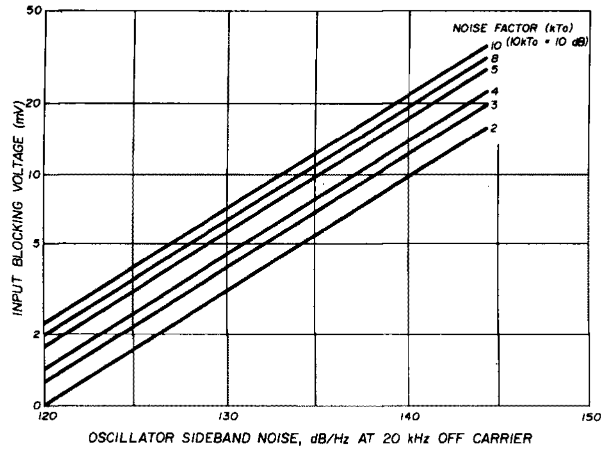


fig. 22. Receiver blocking effect caused by oscillator sideband noise. Blocking can be improved dramatically by ensuring that oscillator sideband noise is suppressed by careful circuit design.

the input transformer of crystal filters using toroids. Two 1-volt input emf signals at 50 ohms must not produce third- and second-order intermodulation distortion products of 85 dB or less. **Fig. 20** shows a typical configuration of a 6-pole vhf crystal filter with high distortion because of the input toroidal transformer, while **fig. 21** shows an improved version using a tuned input transformer with almost no measurable distortion.

Oscillator sideband noise. As explained earlier, oscillator sideband noise contributes significantly to the large-signal-handling capability of a receiver. Blocking and 3-dB compression are often confused in the literature. Compression is the effect caused by sideband noise, i.e., 20 kHz off the vfo carrier frequency, which causes blocking. This blocking effect can be expressed in terms of the sideband noise as it affects receiver sensitivity:

> Assume sideband noise of 145 dB per hertz located 20 kHz from the vfo carrier frequency and a receiver noise figure of 10 dB. An input signal of 50 mV will cause 3-dB blocking or desensitization of the receiver, while 3-dB compression may occur at a point as high as 1 volt on the input signal. This relationship is shown in **fig. 22.**

When designing oscillators with very low sideband noise, either selected fets operating in a saturated mode or medium-power 2N3866 transistors should be used.

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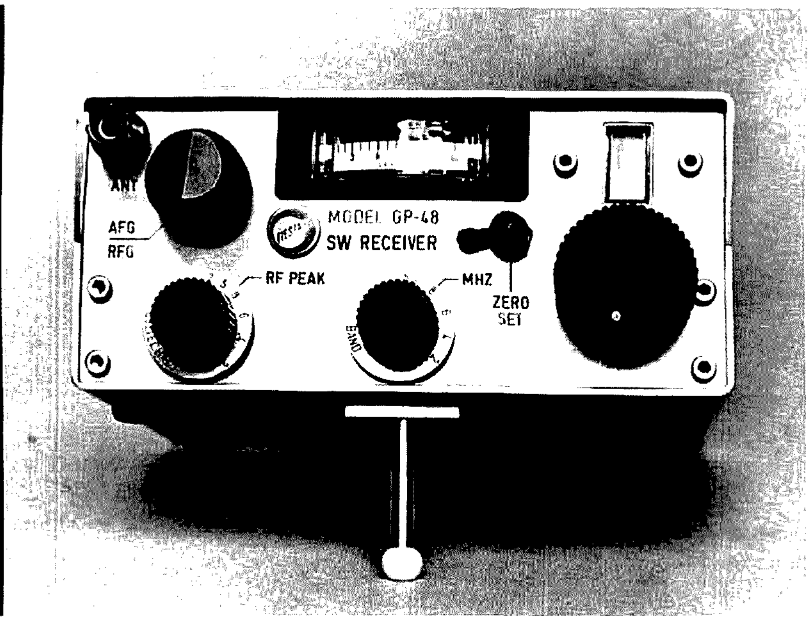


double-conversion hf receiver

with mechanical frequency readout

Although designed for shortwave broadcast, this receiver is easily modified to provide coverage between 3.2 and 30 MHz

Previous articles have described my ideas on monoband versus multiband receivers.[1,2] The concepts developed in these articles are still true, but response to another article[3] indicates many experimenters aren't interested in building a one-band receiver and are willing to accept spurious responses in a multiband unit. This is particularly true for amateur-band operation where it's a simple matter to change frequency if a contact is made where a spurious response exists in the receiver. My operation is largely confined to listening to shortwave broadcast. Tuning a signal on these bands is a different situation. Shortwave broadcast stations emit signals on one frequency. If one of these signals happens to fall on a spurious response in the receiver, reception is difficult and a rare DX station may be lost.

This article presents the results of my experiments to produce a miniaturized, solid-state receiver with a mechanical counter digital frequency readout. The receiver covers 4.2 - 5.2 MHz in five bands, but with modification to front-end inductances and suitable crystal selection, it can provide general coverage between 3.2 and 30 MHz.

conversion method

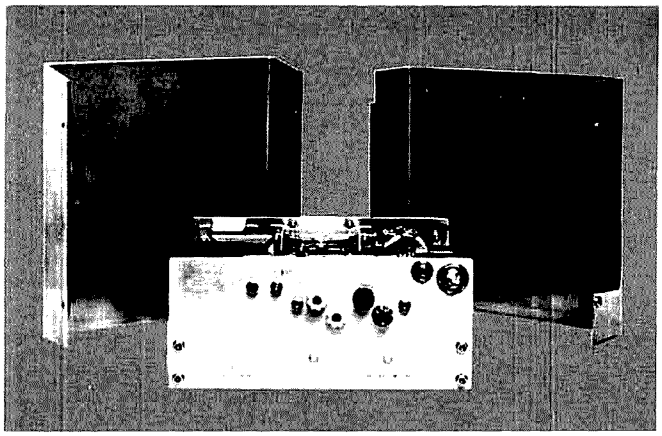
An extensive search was made of commercial and military communications gear. Some mixing combinations, excellent as they are, were discarded because of their complexity. This was particularily true of the Collins 51J4, whose mechanical complexity is high and for the Racal 217 whose electrical complexity is high. Other combinations were also discarded. The Collins 75A4, for instance, would be difficult to duplicate because of the complexity required to build a linear vfo over 1 MHz in the 2-MHz region.

The Collins 75S3 series was screened in depth. I operated a 75S3 inside and outside the amateur bands for several years and found only one spurious response

By Jack Perolo, PY2EIC, P.O. Box 2390, San Paulo, Brazil

at 5000 MHz caused by the second harmonic of the vfo when tuned to 2500 MHz. This occurred when 5 MHz was tuned in the 4.8-5.0 MHz band; by switching to the 5.0-5.2 MHz band, 5000 MHz was received with no problem.

Among the advantages a duplication of the Collins 75S3 would offer is its straightforward construction. And, even more important by amateur standards, the vfo tunes only 200 kHz making it easy to linearize. The price you have to pay for these advantages, however, is that if the vfo frequency excursion is only 200 kHz, each high-frequency crystal provides coverage for a band 200-kHz wide. In other words, if you wish to use the receiver over a wide frequency range, the quantity of crystals required is somewhat high. The model GP48



Rear view showning stainless-steel cabinet construction details together with the back-panel 110-Vac input (white jacks) and 6-Vdc input (dark jacks). Two phone jacks in parallel allow headphone operation with standard and miniature plugs. This receiver required about $150 worth of parts and 150-200 hours construction time.

receiver described here has only five bands and therefore five crystals; earlier models had wider coverage: the GP45 had ten bands and the GP46 36 bands. At about $4 per crystal, the GP46 had about $150 worth of crystals.

I built some solid-state receivers around the Collins 75S3 mixer combination. I was quite pleased with these experimental efforts; then I made another unit (dubbed GP42). The GP48 is the sixth of a series. Its schematic appears in fig. 2. This receiver covers 4.2-5.2 MHz or, generally speaking, the 60-meter shortwave broadcast band. With suitable crystal selection and front-end coil inductances, any frequency between 3.2 and 30 MHz can be covered with the exception of 5.2-6.5 MHz.

circuit considerations

The basic circuit is closely related to the receivers in references 1 and 3, whereas the basic mixing scheme is derived from the Collins 75S3 line. There are, however, some differences worthy of mention. The Collins front end is slug tuned, with parallel capacitors switched in to

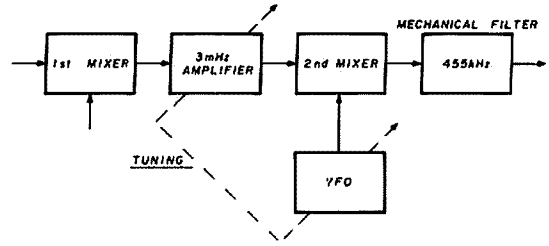


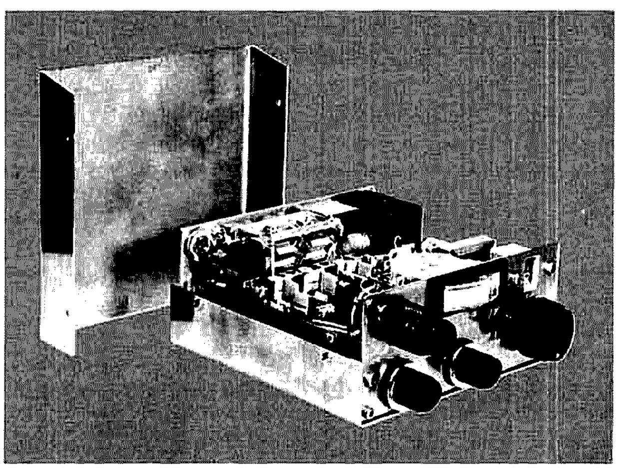
fig. 1. Block diagram or receiver 3.MHz i-f strip showing tuning method.

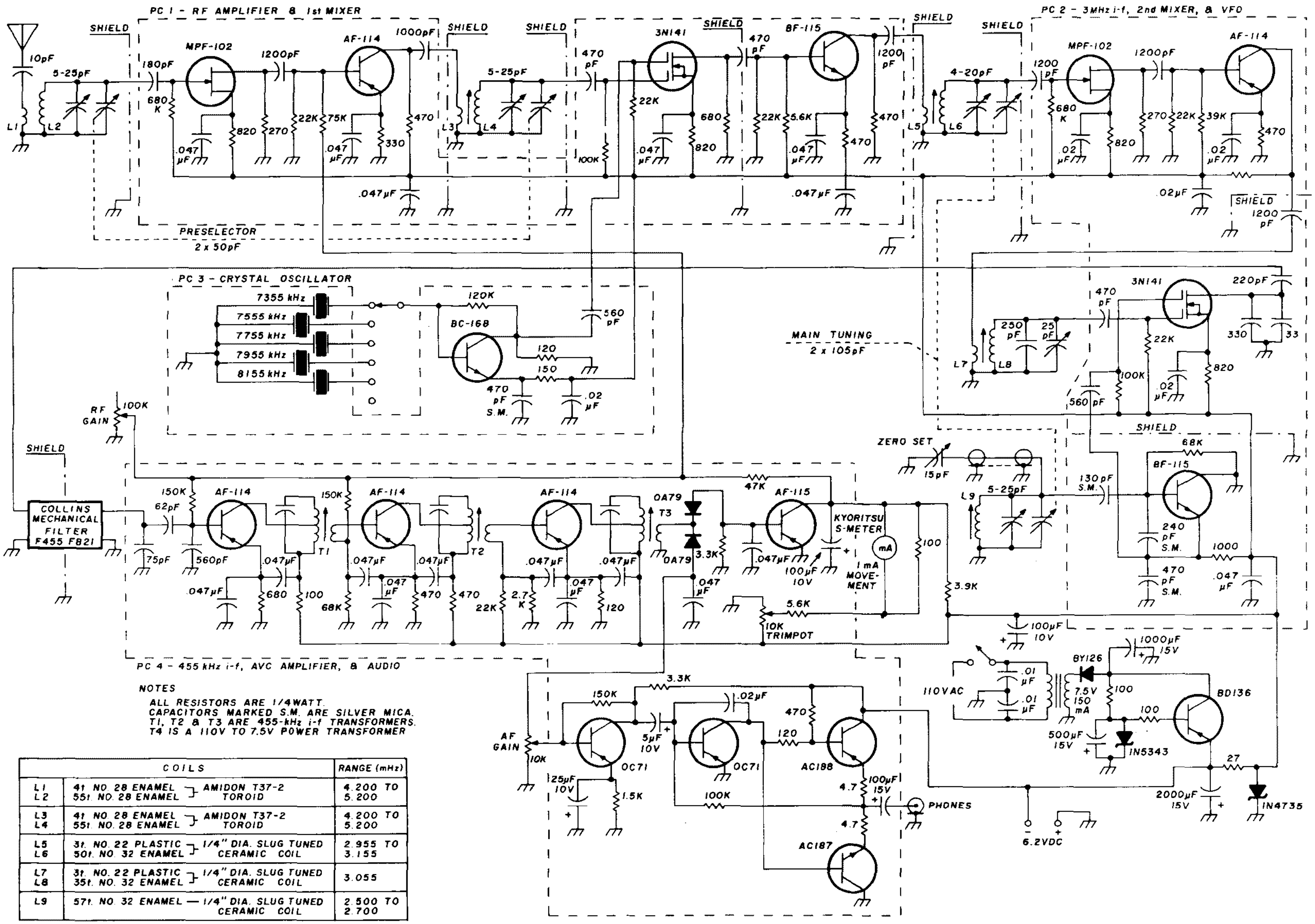
cover the low-frequency bands. While I've built similar units, I preferred the usual variable capacitor combination for this project, as most builders would probably not be able to homebrew a backlash-free multicore mechanism. For those who prefer the permeability tuned version, some surplus outlets offer powdered-iron cores with high length-to-diameter ratios and with a spring (rather than the usual threaded pin) termination. The permeability, μ, of the cores is unknown, but it could be found with a grid dip meter or by breadboarding the front end before final assembly.

The Collins 3-MHz i-f is fixed tuned with a remarkably flat response over the 2955-3155 kHz range. I didn't wish to compromise on sensitivity and, after some experimenting, I decided to make the 3-MHz i-f partially tunable (fig. 1). The vfo capacitor is a two-gang affair that tunes the vfo and the input to the 3-MHz i-f amplifier, whereas the second mixer is fixed tuned. This improves the bandpass flatness and, by having a fixed-tuned mixer, buffer stages aren't needed between the vfo and the mixer. In fact, since the mixer is fixed tuned and operates over a narrow frequency range (200 kHz), it offers a reasonably constant load to the vfo.

This brings up another difference between my design and that of the Collins 75S3 line; the vfo is capacitance-tuned in my case. I have already published an article on

View with upper half of cabinet removed. Front panel controls are (from left) antenna input jack, af/rf gain control, preselector tuning control, band switch, zero-frequency set control, and main-tuning control.



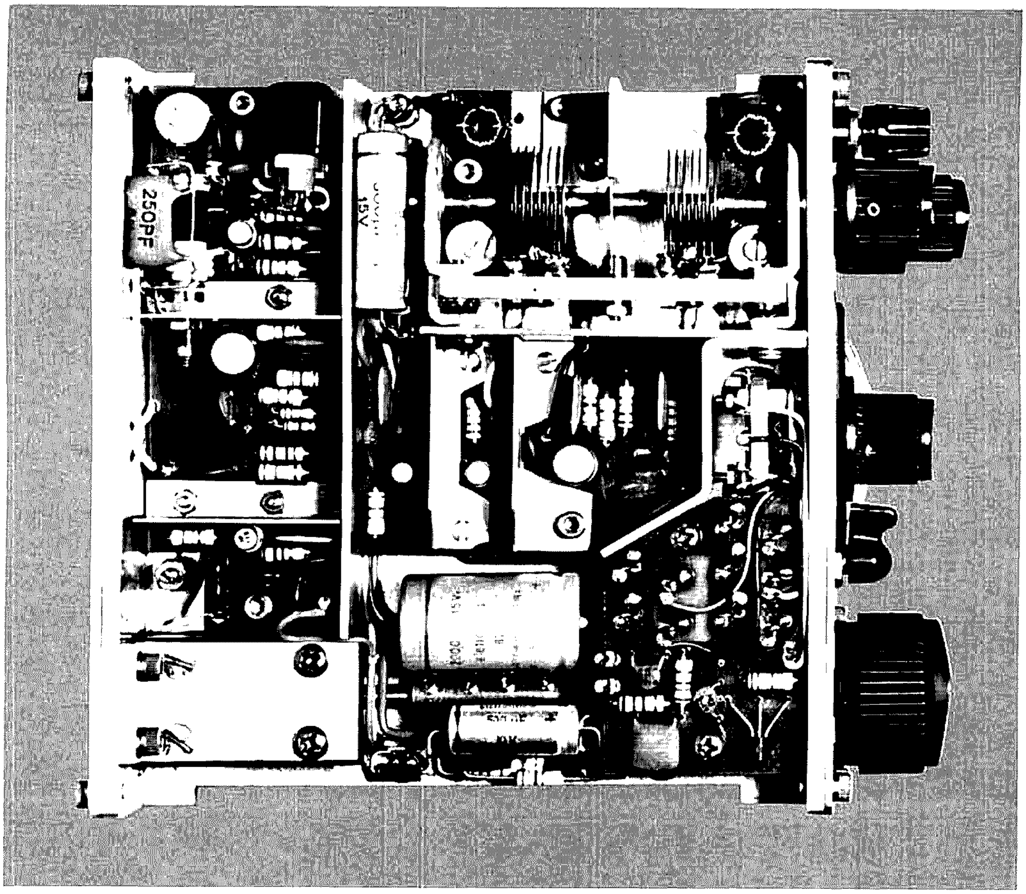


NOTES

ALL RESISTORS ARE 1/4 WATT.
CAPACITORS MARKED S.M. ARE SILVER MICA.
T1, T2 & T3 ARE 455-kHz i-f TRANSFORMERS.
T4 IS A 110V TO 7.5V POWER TRANSFORMER

	COILS	RANGE (mHz)
L1 L2	4t NO. 28 ENAMEL 55t. NO. 28 ENAMEL — AMIDON T37-2 TOROID	4.200 TO 5.200
L3 L4	4t NO. 28 ENAMEL 55t. NO. 28 ENAMEL — AMIDON T37-2 TOROID	4.200 TO 5.200
L5 L6	3t. NO. 22 PLASTIC 50t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	2.955 TO 3.155
L7 L8	3t. NO. 22 PLASTIC 35t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	3.055
L9	57t. NO. 32 ENAMEL — 1/4" DIA. SLUG TUNED CERAMIC COIL	2.500 TO 2.700

fig. 2. Receiver schematic. Crystal frequencies shown are for receiver coverage between 4.2-5.2 MHz. Receiver range may be changed to include 3.2-30 MHz by substitution of suitable crystals and front-end inductances.



Bottom view. Mechanical filter is at top left next to the mixer coil and its circuit. The 3-MHz two-stage i-f amplifier is at left center; the vfo circuit is at the bottom next to the gear reducer. At top right is the two-gang front-end variable capacitor with its toroids and trimmers. The rf and mixer circuits are at center; the band switch and crystal oscillator are at right (crystals are under the PC board). At bottom center is the regulated power supply.

a permeability tuned vfo[4] as well as some generalized considerations on linear vfos.[5] Again, the average amateur will find if difficult to come up with a backlash-free permeability-tuned vfo and even more difficult, as in this case, to gang two slugs together.

construction details

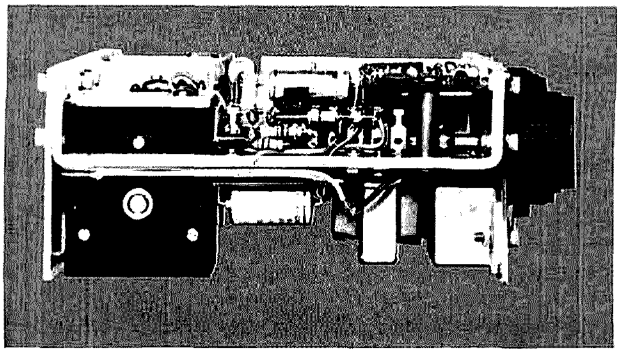
The photos show different views of the GP48; most of the construction details can be seen in these pictures. The receiver dimensions are 6 inches wide by 2.5 inches high by 6 inches deep (15 by 6 by 15 cm). All front and back panel lettering was made with a pantograph; knob skirts were made from 16-gauge stainless steel and were similarly engraved. Aluminum shields are 1/32 inch (0.8mm) for PC-board mounting and 1/16 inch (1.6mm) for chassis mounting. Extensive shielding and capacitive decoupling assure stable operation at high gain. High-quality components are used throughout.

The gear-reduction unit, which has a 100:1 ratio, is permanently lubricated in a sealed container. It's a British import made by Muffett, Ltd. The counter, made by Veeder Root, has three vertical digits. A vertical counter eliminates parallax problems when the receiver is tilted upward. The receiver tuning rate is 10 kHz per knob revolution, which makes for exceptionally smooth action. The brass bevel gears for the counter, made by Boston Gears, have a 1:1 ratio.

The vfo double-gang variable capacitor is also a British import, available in the U.S. from the J.W. Miller Company. The capacitor has very low torque, with ball bearings at both ends. It is coupled by a flexible (bellows) joint to the worm gear reducer. The S-meter is a Japanese import with a 1-mA movement. The meter face was replaced with one reading in S-units.

The receiver cabinet, which is in two halves, is made of 18-gauge stainless steel, secured by 1/8 by 3/4 by 4½ inch (3 by 19 by 114 mm) spacer bars and four 4-40 (M3) binder-head screws. Four glass-epoxy PC boards are used. The i-f/af board is mounted on top of the chassis, while the vfo/3-MHz i-f/second-mixer, together with the front-end rf amplifier/first mixer and crystal oscillator/crystal holder, are mounted below chassis.

The vfo circuit was derived from previous receivers and its circuit adapted to cover the 2500-2700 kHz



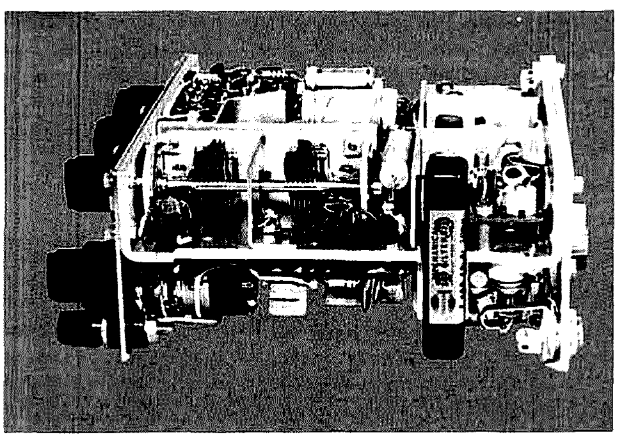
Right side. Gear reducer is at left; regulated power supply at center top. At top right is the crystal oscillator PC board with the crystals barely visible below. The power transformer is at the bottom right next to the mechanical counter. Both front and back panels are fastened with 6-32 (M 3/S) Allen-head stainless steel screws.

range; the variable capacitor (105 pF) is correct for the frequency excursion in question. The audio strip ends with a complementary pair wired to avoid transformers. Output power exceeds 0.5 W – more than enough for headphones. The power supply is electronically regulated providing, through back panel jacks, the dc power to run other gear, or conversely, to be used to feed the receiver from a dc source.

circuit details

Because of the extremely high gain of this set, proper shielding and bypassing must be used to achieve stability. All tuned circuits are shielded from the circuits of the corresponding active devices. Heavy filtering is used on all power lines. The vfo coil is shielded to minimize proximity effect. The vfo capacitor is not shielded because it must be accessible to allow for easy filing of its plates during the linearization procedure (described below). It must be remembered, however, that during the filing process, a small temporary shield must be fastened to the top of the vfo variable capacitor to

Left side. Front-end double-gang variable capacitor, toroids, and trimmers are at upper left, installed on small printed circuit boards.



simulate the effect of the metal receiver cabinet. Failure to do so will result in degrading vfo linearity because of the proximity of the cabinet when set in place.

It is important to understand that when a vfo is coupled to a counter, a simple change such as moving a wiring harness can degrade the calibration by more than one kHz at the band edge. Therefore, before starting to linearize the vfo capacitor all screws, nuts and components in general must be securely tightened in place and no circuit changes should be made during or after linearization. The easiest way to achieve perfect linearity is to file down the capacitor plates to within 1 kHz or so of the nominal frequency desired; from then on, the fine part of the linearization procedure is to slightly bend and adjust the side plates of the capacitor rotor. This procedure has the advantage of being reversible whereas filing is not.

selectivity improvements

A major difference between amateur-band operation and shortwave broadcast listening is in selectivity. In the

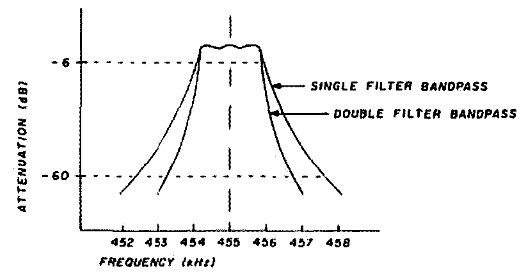
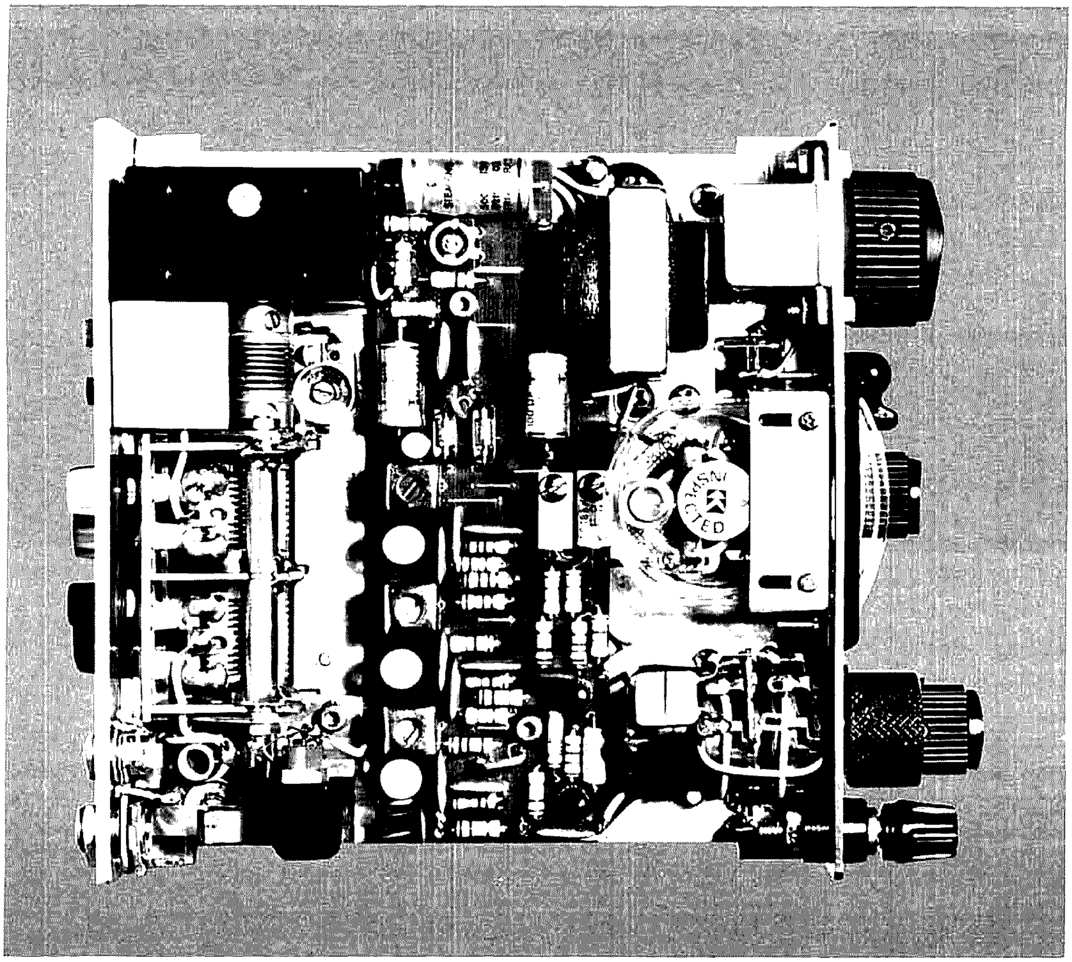


fig. 3. Skirt response of single and double filter arrangements. When cascaded, each filter must have an identical midband frequency for best results.

first case interference may be generally avoided by changing frequency; obviously this isn't possible when receiving shortwave broadcast stations. The minimum bandwidth for a-m reception is stated in the literature as 3 kHz or thereabouts. Any serious listener knows this number is way too high. I've used a 2.1-kHz bandwidth (through a mechanical filter) for many years but have changed to narrower bandwidths as conditions demand. For serious shortwave broadcast listening, I'd say the ideal receiver bandwidth is 1.8 kHz.

While mechanical filters are available with a skirt ratio of 2:1 (-60 to -6 dB), a nearby station can still cause interference because of the bell-shaped response of these filters. A remarkable improvement can be achieved by cascading two identical filters, but some words of caution are in order. Using this method, the skirt response at -60 dB can be improved by a factor of $\sqrt{2}$, or about 1.4 times, while maintaining the response constant at -6 dB (see fig. 3). A definite improvement in obtained; so much in fact that comparing two receivers tuned to the same station, one with a 1.4-kHz mechanical filter and the other with two cascaded 2.1-kHz mechanical filters (both bandpass figures at -6 dB), one's reaction is that the second receiver is more selective than the first.

To realize the full potential of this combination it's



Top view. Gear reducer is at top left. One of the 3-MHz i-f coils is visible below the variable capacitor. PC board in center includes the three i-f amplifiers, avc amplifier, and audio strip. At top right are the power transformer, mechanical counter, S-meter, rf/af control, and antenna input jack. Aluminum bracket, top left, shields the vfo coil to minimize proximity effect when the steel cabinet is installed.

imperative to use two filters of identical midband frequency to avoid stagger tuning with consequent skirt degradation. Each filter must be individually shielded to avoid ground loops and to ensure the signal travels through the filters and not around them. You can't take excessive precautions in this respect, because such arrangements are critical beyond imagination. The GP46 receiver (mentioned earlier) was based on this concept with entirely satisfactory results. Another i-f stage should be added to compensate for the insertion loss of the second filter.

frequency readout

While an electronic frequency readout could be used, a mechanical counter is less expensive and physically smaller. For portable work, the mechanical counter is also better since it requires no power. I believe that the vfo linearization work, however tedious and delicate, is still more advantageous than building an electronic counter. Such a counter, however, could always be added to the receiver as an external unit. The advantage in this case is that the counter can be used for other projects. I've found that an external counter offers less interference than built-in units, as their 100-kHz clocks tend to show some leakage into the receiver circuits.

acknowledgement

I'd like to thank PY2GP for his continued support in the realization of this and other projects.

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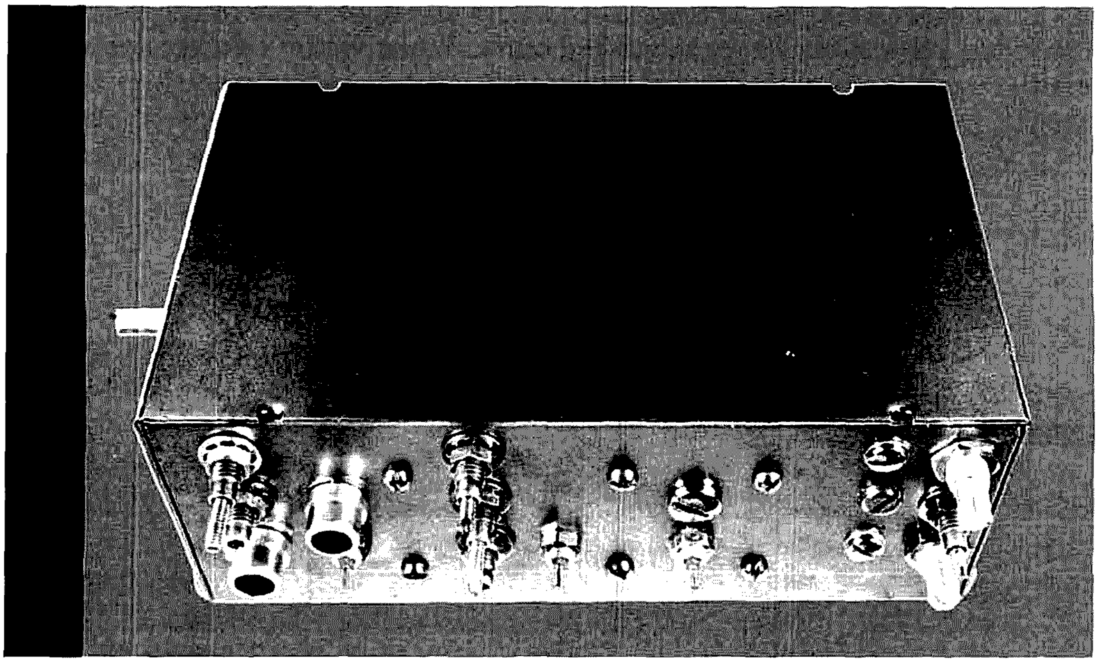


multiband high-frequency converter

A VVC-tuned converter that extends your 80-meter receiver tuning range to include 40-10 meters plus WWV

Many receiver designs have been published for the 80-meter band. This article provides complete design and construction details for a multiple-band converter to extend the range of this type of receiver. This design provides conversion of the hf bands and 10-MHz WWV signals to 3.4-4.0 MHz for i-f amplification and detection.

Fig. 1 illustrates the converter design in which mos field-effect transistors are used in the rf amplifier and mixer stages because of their superior spurious-response rejection and signal-handling capability. The dual-gate protected mosfet features rf gain or agc control in the rf-amplifier (Q1) stage and outstanding mixing characteristics for conversion to the i-f output (Q2).

Overall gain of the converter is shown in table 1. Coil Q for each stage is fairly high; as a result bandwidths are narrow, so some method of tuning is necessary to cover each band. Tuning is by variable capacitance (VVC) diodes in each of the rf and mixer tuned circuits.

circuit description

Fig. 2 is a schematic of the rf and mixer stages. Selection of each band is by band switching the appro-

By M.A. Chapman, K6SDX, 935 Elmview Drive, Encinitas, California 92024

priate tuned circuit in the rf and mixer input stages. The mixer output uses a VVC that tunes a full MHz. The individual tuned circuit design centers are for the middle of each amateur band. The VVC will allow the coils to cover each band with a margin of several hundred kHz at the band edges. The unloaded Q of the mixer output is approximately 150; the bandwidth is quite sharp and will require simultaneous tuning with the rf stages.

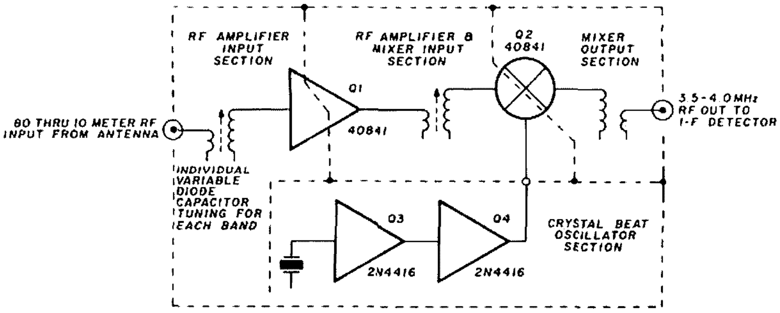


fig. 1. Converter block diagram. Design covers 80-10 meters and 10 MHz for WWV.

The total converter bandwidth is difficult to specify because of the variety of interpretations. The half-power bandwidth – that is, the width for 50% decrease in converted signal amplitude – was measured at less than 200 kHz on all bands. This sharp selectivity also indicates that for a typical 500-kHz band-segment width, several peaking adjustments are required from end to end. Both the rf and mixer stages depend on source resistance for gate bias. The source bias in Q2 is extremely important since it establishes the transfer curve linearity that provides an optimum combination of mixing and spurious response rejection (more information about this source bias and mixer linearity is given in the mixer discussion that follows).

Converter interior. Shielded partitions contain (l to r) oscillator, mixer output, rf amplifier output, and rf amplifier input sections.

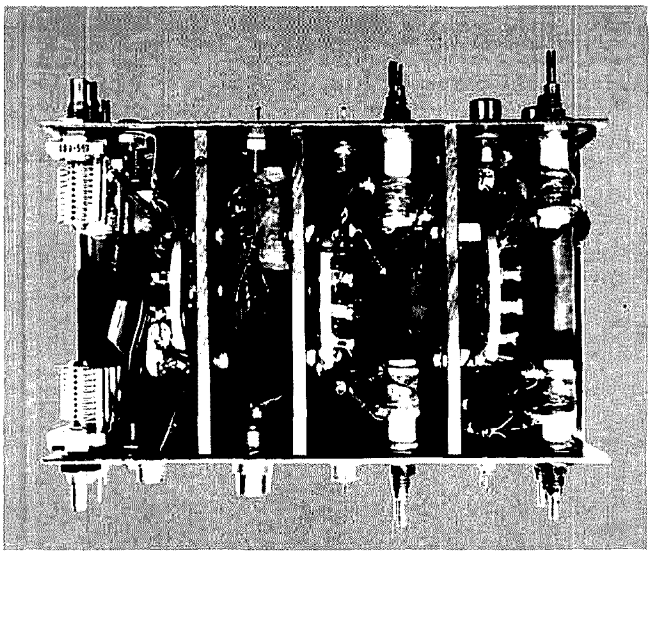


Fig. 3 is the high frequency beat oscillator schematic. This circuit is an adaptation of those in references 1 and 2. A conventional LC network is illustrated for 10-meter beat-oscillator conversion. This signal could have been generated by a crystal as for the other bands; however, fundamental crystals above 20 MHz are expensive and restrict the circuit to some fixed segment of the 10-meter band. Down conversion is accomplished for all bands using a beat oscillator signal approximately 3.5 MHz less than the lower edge of the receiving frequency. For 7.0-7.3 MHz conversion we use the top end of the 80-meter band, 3.7-4.0 MHz, as the i-f output to minimize beat-oscillator feedthrough in the mixer output stage. The 20-meter band (14.0-14.25 MHz) is down converted to the middle of the 80-meter i-f range to maximize circuit selectivity in subsequent stages.

table 1. Performance summary

band	tuning range (MHz)	i-f output (MHz)	converter gain (dB)
80M	3.5-4.0	3.5-4.0	0
40M	7-7.3	3.7-4.0	37
20M	14-14.25	3.625-3.875	41.7
15M	21-21.45	3.5-3.95	36.5
10M	28.5-30 (any 500 kHz segment)	3.5-4.0	34.5
WWV	9.9-10.1	3.65-3.85	28.9

Sensitivity: <.15V rms using the i-f system described in reference 3 on all bands.
Bandwidth: ±100 kHz for 50% decrease in signal amplitude without peaking adjustment.
Spurious signal rejection: ≈50 dB attenuation at ±1 MHz.

Regular 32-40 pF, parallel-tuned, fundamental-

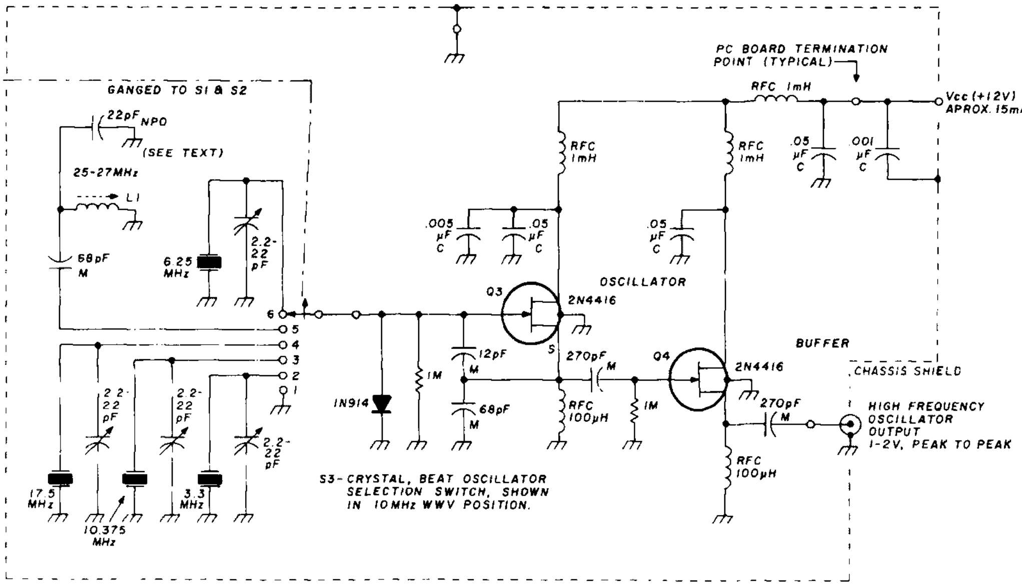


fig. 3. Oscillator schematic. L1 consists of 11 turns of AWG 26 (0.3mm) enameled wire on J.W. Miller 4500-2 form.

frequency crystals should be used. The 2.2-22 pF parallel trimmer capacitors across the crystals may be deleted if you're willing to accept some error in the converted signal output. This error in most cases will be less than ±5 kHz if the capacitors are not used. The total capacitance across the crystal is the sum of the parallel component capacitance and the gate source capacitance. Since all of these components contain variations in capacitance tolerance, some compensation or trimming is required for exact beat-oscillator frequency generation.

Where digital readout schemes are used for frequency display and consideration has been given for band-edge

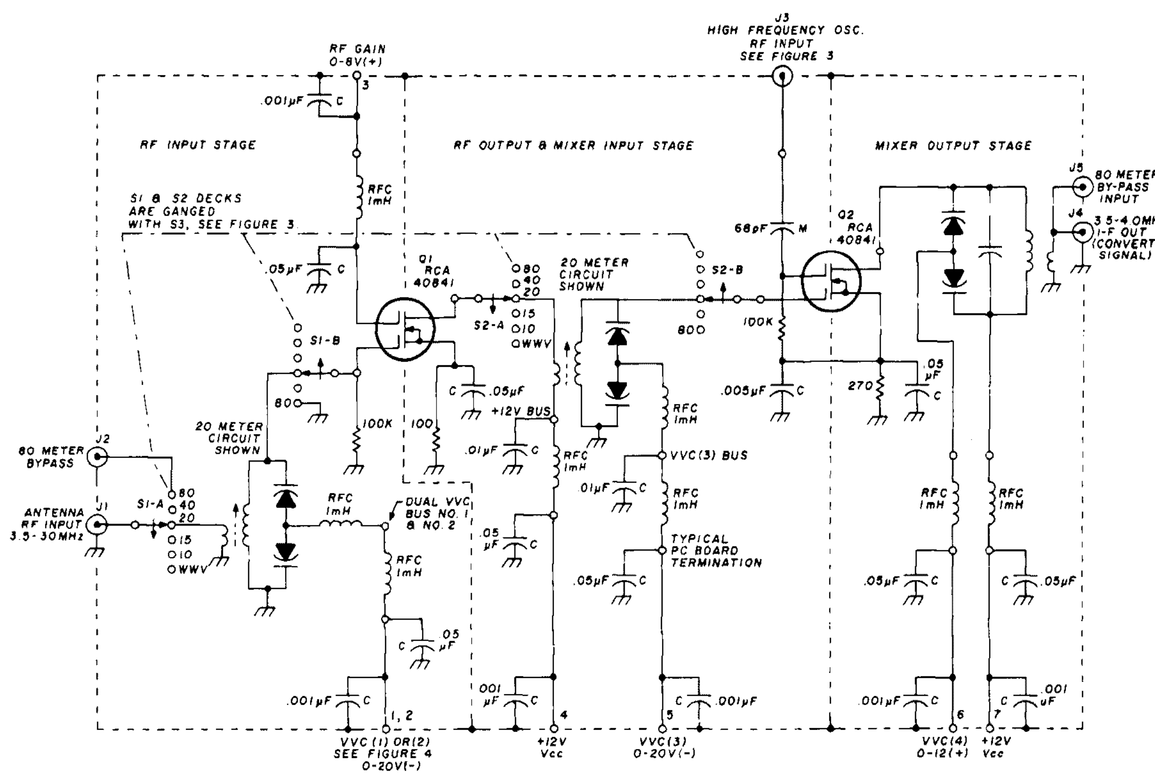


fig. 2. Rf amplifier and mixer schematic. Variable capacitance diode tuning is used in each stage.

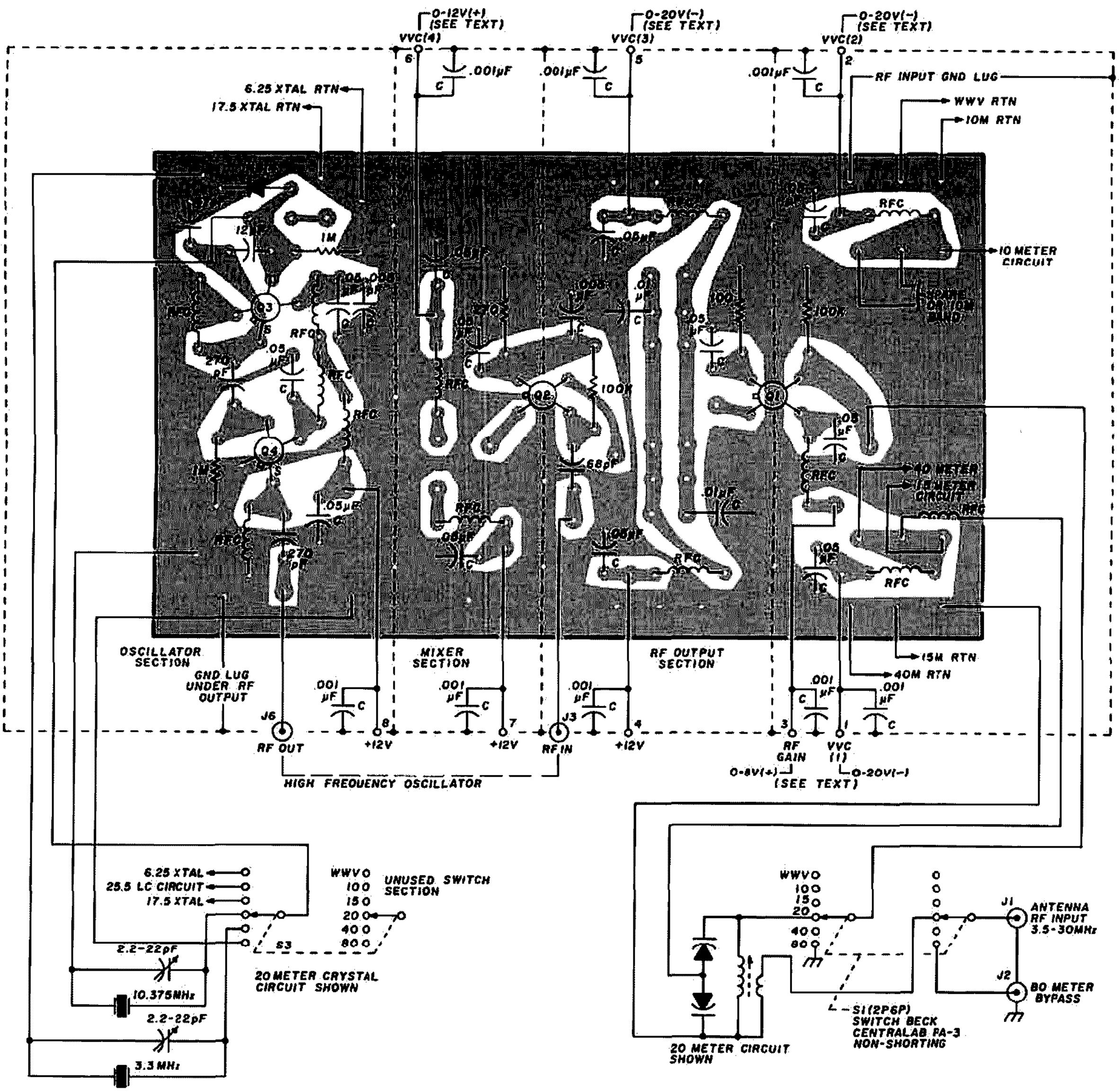


fig. 4. PC-board component layout for rf amplifier and oscillator.

determination, this frequency error is continuously known. If you depend on a slide rule dial visual display of the tuned frequency, then the oscillator crystal frequencies should be set precisely using some type of capacitor trimming similar to that indicated in the diagram. Both oscillator and buffer use a source inductive load. Considering the wide range of oscillator frequencies, this method of buffering the oscillator produces a stable, moderately uniform output level and minimizes spurious frequencies, which could fall within the tuned frequency range. In combination, the rf mixer and oscillator stages are joined to provide a comfortable 300-500 kHz segment from each of the hf bands to be converted to the 3.5-4.0 MHz 80-meter band. Because of the wide tuning capability in the rf and mixer stages, MARS and general-purpose tuning can also be added if sufficient i-f tuning range is available.

construction

From the photos it's seen that the converter is built around a simple PC board* with attached aluminum plates acting as interstage shields and a convenient mount for the switch decks. This PC board with the attached shield plates is mounted in a standard LMB 136 aluminum chassis box. The shield plates are aligned by drilling the switch shaft clearance hole through all three plates at the same time, then centering one plate on the

*Undrilled PC boards are available from the author for $3.00 including postage.

end of the box to locate and align the box and shaft clearance hole. The PC board is attached to the bottom edge of the plates using self-tapping screws. The shield plates are fastened to the sides of the box in the same manner. This construction follows the same general approach described in reference 3.

To reduce spurious noise and birdies in other portions of the receiver, generous use is made of bypass capacitors and rf chokes both at the PC board and at the point on the chassis side where control and supply voltages are applied. Each stage is individually separated and control lines decoupled from each other to minimize feedback. An examination of figs. 4 and 5 shows potential bandswitching of six individual bands; however, only five are used in this design because of physical limitations of the box and coils. By using a slightly larger box, an additional portion of the 10-meter band could be included or perhaps a 6-meter segment.

Not everyone will be interested in the conversion of all the bands shown. For instance, if only two or three bands are desired, considerably more room would be available for access. If only one band conversion is desired, adequate room is available to mount the coil components directly to the PC-board groundplane areas. For those who may wish to experiment with a less complex and expensive system of fixed-band conversion, the capacitor resonant values of fig. 6 may be used. Padding the coils with a 10k, ½-watt resistor will decrease coil Q sufficiently so that no rf tuning would be required over most of the 500-kHz band segment. The VVC control wouldn't be required, resulting in a further reduction in power-supply voltages.

The PC layout of fig. 4 also indicates an unused pad next to Q3 gate. The purpose of this pad is for the installation of a coupling capacitor between the oscillator switch deck common and Q3 gate to allow for conventional LC networks to be used for all oscillator frequencies instead of crystals, similar to the method for generating high-frequency oscillator signals for 10 meters as in fig. 3.

oscillator

Construction should be accomplished one stage at a time. The easiest place to start is with the crystal oscillator. It's not necessary to have the shields or switch decks in place to verify crystal-oscillator operation. Mount the oscillator components to the PC board as indicated in fig. 4. Temporarily install any fundamental frequency crystal between 3-30 MHz between the PC pad from Q3 gate and the ground plane where holes are provided. Install a 100 ohm current-limiting resistor from the +12V PC pad and apply power. The crystal signal should appear at the source of both Q3 and at the output coupling capacitor (270 pF) from the Q4 source.

mixer

The mixer output stage should be wired next. Estimate the service lead length required from the output coil to the board, and allow the coil to hang free from its leads. The bandswitch and shield between the input and output aren't required for initial testing. Install 100-ohm current-limiting resistors to the +12V mixer coil PC pad and the VVC (4) control PC bus (0-12V). By applying a low-level 3.5-4.0 MHz signal to Q2 gate 1 and varying

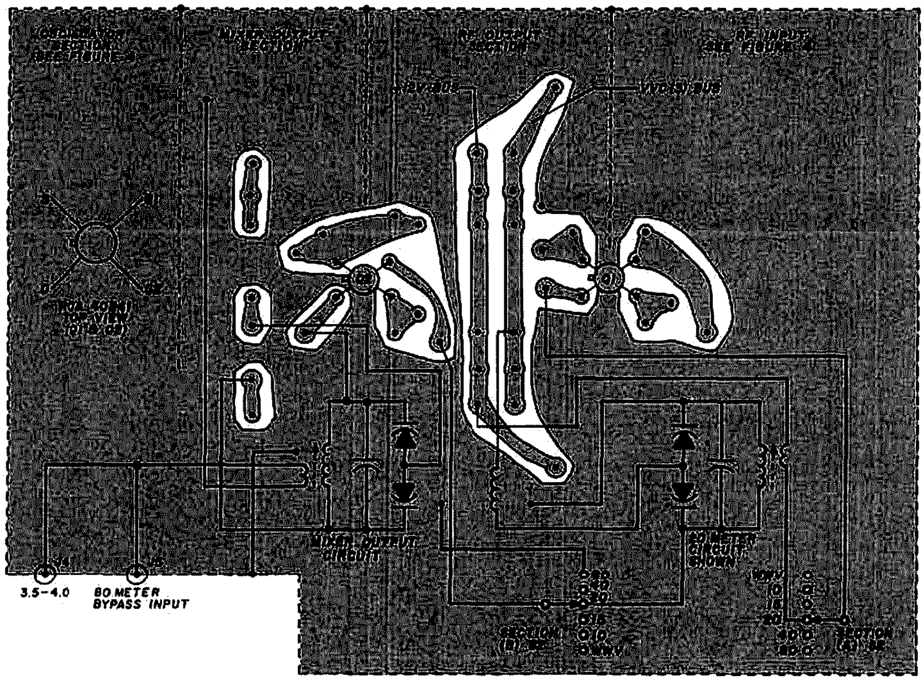
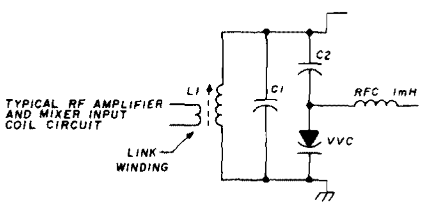


fig. 5. Rf and mixer tuned-circuit interconnections.

the VVC voltage on the output coil, signal gain and selection should be apparent. The network will be detuned because of oscilloscope internal shunt capacitance if a scope probe is not placed on the coil link output. In this mode we're using Q2 as a simple single-channel fet amplifier, tuning the drain circuit and monitoring the link. Additional verification can be obtained by temporarily grounding Q2 gate 2; the output signal should fall off immediately. To simplify these observations the signal applied to gate 1 should be modulated so that a low-frequency audio envelope can be monitored by an oscilloscope. To verify Q2 mixing characteristics and to optimize the circuit, substitute a 500-ohm potentiometer in place of Q2 source resistor. Add a jumper wire from the oscillator output PC pad (Q4 source capacitor) to the input coupling capacitor on Q2 gate 2.

table 2. Rf tuned-circuit component complement



band/ freq	L1 form[4]	Q[2]	rf amplifier coils winding[1] turns	AWG	(mm)	link[1] turns	AWG	(mm)	nominal cap (pF)	C1 (pF)	C2[3] (pF)	VVC type
40m[3]	4500-2	65	25	28	(0.3)	5	30	(0.25)	175	22	—	MV1666(2)
20m[3]	4500-3	80	20	28	(0.3)	5	30	(0.25)	65	—	—	MV1652(2)
15m	4500-3	65	13	26	(0.3)	4	28	(0.3)	58	—	68	MV1660
10m	4500-6	60	10	26	(0.3)	3.5	28	(0.3)	45	—	43	MV1660
10MHz	4500-2	60	23	28	(0.3)	4	30	(0.25)	82	82	—	—

Notes:
1. Turns are close wound, slightly loose over form.
2. Unloaded value.
3. C2 is a VVC, mounted anode-to-anode. Q dope all components after soldering.
4. D.W. Miller part numbers.

freq (MHz)	L1 form	Q[2]	mixer coil winding[1] turns	AWG	(mm)	link[1] turns	AWG	(mm)	nominal cap (pF)	C1 (pF)	C2 (pF)
3.5/4.0	plex rod 3/8 in. (9.5mm)dia	90	48	28	(0.3)	10	30	(0.25)	185-245	150	MV1403(2) VVC mounted anode-to-anode

Notes:
1. Winding is 3/4 in. (2cm) long located along center of rod. Link is on bottom end near chassis.
2. Unloaded value.

Any of the indicated crystals may be used in the oscillator; however, let's assume that the 10.375-MHz crystal is used and temporarily mounted in the PC board oscillator section as previously discussed. By applying a low-level modulated 14-MHz signal to Q2 gate 1 with the beat oscillator signal applied to gate 2, and by adjusting the VVC (4) control voltage on the mixer coil, a signal envelope similar to that shown in **fig. 7** should be seen while monitoring the mixer output coil link. Mixing linearity may be optimized by adjusting the temporarily installed 500-ohm potentiometer for both mixer gain and peak signal-to-local oscillator feedthrough ratios. Next, remove the potentiometer and measure its resistance. Select a fixed resistor of this value, which will be between 200 and 300 ohms. This resistor establishes the gate 1-to-source bias on Q2 for linearity of the fet transfer characteristics. For ideal mixing the transconductance curve should approach a straight line when the drain current is plotted against the gate 1 voltage. Since the zero gate voltage versus drain current varies over a wide margin, selection of this source resistor will optimize the device operation.

rf amplifier

The rf amplifier PC components should be installed next. As in the previous steps, no interstage shield is necessary for initial testing. From the pad on the PC board for Q1 drain circuit, temporarily install a 0.01μF capacitor to the gate 1 pad of Q2 and add a 100-ohm resistor between the Q1 drain PC pad and the +12V bus. Apply power to the rf and mixer drain circuits and oscillator, and apply a low-level modulated 14-MHz signal to Q1 gate 1. The mixer output should be similar to the previously described mixer test signal, with some high-frequency feedthrough apparent. Q1 gain control can be verified by temporarily grounding Q1 gate 2; the output signal should fall off immediately then slowly increase when gate 2 is left open. The rf and agc gain control features of a dual-gate mosfet can be observed if a variable positive voltage is applied to Q1 gate 2. The output signal should be nearly zero when gate 2 is at or near ground potential; then the output signal will rise in amplitude as the gate 2 voltage is increased to about 8 Vdc. Maximum gain is achieved when the gate 2-to-source voltage is about 6 Vdc.

You may want to adjust the Q1 source resistor. Because Q1 is a depletion device, the source could be operated at ground potential with maximum gain from the stage; however, some biasing of gate 1 is desirable for device stability, and 100 ohms is a common value for a mosfet in this type of rf application. For higher frequencies, such as 144 or 220 MHz, a 220-ohm resistor would

provide some improvement in the noise figure with a sacrifice in gain. If you need the extra gain, the source resistor could be reduced to 47 ohms. If plenty of selectivity and gain are available in the i-f amplifier, my suggestion would be to increase Q1 source bias to several hundred ohms.

final adjustment

At this point we're confident of our component operation, and all temporary connections should be removed. Figs. 4 and 5 illustrate the inductor and switch deck interconnection. The switches select the crystal beat oscillator frequency and appropriate tuning network for the desired band. In the 80-meter position the signal bypasses the rf mixer stages and is presented directly to the output connector. Using 24 AWG (0.5mm) solid hookup wire, adequate strength is available to support the coil components until you're ready to install the PC board into the box and fasten the coil to the chassis sides.

Before installing the PC board and coil assemblies into the box, the interface connections should be temporarily installed as in fig. 7 and a low-level signal applied to each band. This will allow you to de-bug any wiring errors and initially adjust the coil slugs. A signal level of one or two millivolts may be required, depending on your oscilloscope sensitivity, for alignment. A typical modulated 14-MHz signal is shown in fig. 6 and indicates the envelope response with proper tuning and normal gain settings. The signal amplitudes and features may differ considerably if the mixer output is left unloaded. During my initial testing, a 51-ohm, 1-watt resistor was used across the oscilloscope input terminals, and the normal input capacitance was 40 pF.

additional suggestions

VVC selection was based on what was readily available. Nominal resonance capacitance for middle of the band tuning of each coil is shown in table 2. An infinite number of VVC and capacitor values will tune the circuits. The total tuning capacitance of these VVC devices is not fully used for the bands indicated. The tuning ratio for the rf and mixer input VVC devices is approximately 3, and the mixer output VVC device has a tuning ratio of 10. The reader is referred to Motorola's application note[4] for design options.

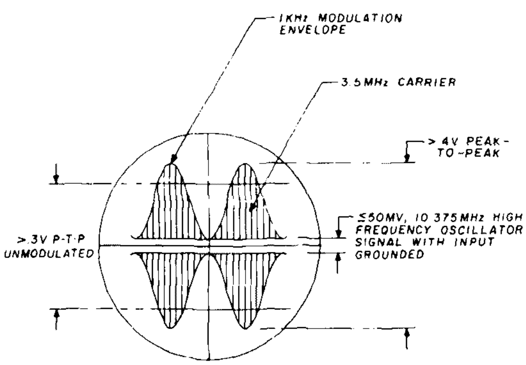


fig. 6. Typical oscilloscope patterns from mixer output with a 1000-microvolt input signal at 14 MHz and 50-ohm resistive load across the scope terminals. Gain is about 40 dB.

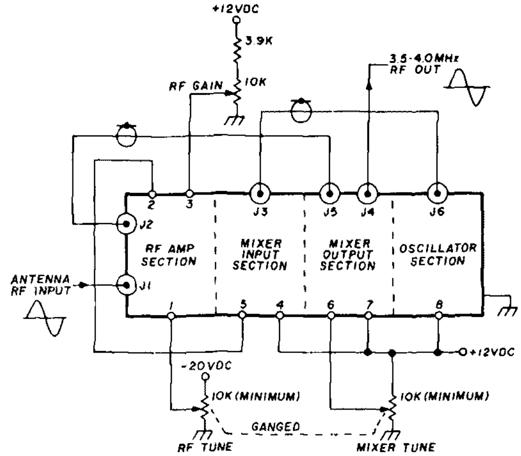


fig. 7. Application diagram. Agc control may be applied to pin 3 of the rf amplifier instead of the rf gain control (see reference 3).

Resistors for the oscillator are ¼-watt; those for the rf and mixer are ½-watt. Low-voltage ceramic or mica capacitors are used. You might experiment using 100-ohm resistors in place of rf chokes for cost reduction; however, the tradeoff here is added spurious noise.

application

Fig. 7 shows the typical interface for VVC control and power bus lines. The rf and mixer tuning controls may be ganged as indicated by adjusting the mixer tuning control for maximum signal output at 3.75 MHz then readjusting the coil slug to the center positions. Tracking between rf and mixer stages is not perfect because of device nonlinearity and frequency tuning ranges; however, for nominal 500 kHz bandwidths, these variations are minimal.

Reference 3 includes an agc circuit that may be used instead of the rf-amplifier gain control. Because of the high input impedance of Q1 control gate, several agc stages may be ganged in parallel; however, the device types in the rf and i-f stages should be similar because of variations in gate 2 voltage and gain between device types. For optimum receiver performance, the VVC controls should remain independent so that signal peaking will be maximum.

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ham radio

solid-state microwave amplifier design

A complete design approach for uhf and microwave amplifiers, and the performance tradeoffs which result with simplified design methods

In two recent articles I introduced a rather simplistic design method for matching microwave transistors in microstripline preamplifiers for 1296 MHz.[1,2] The designs proved entirely acceptable for amateur applications, exhibiting input and output vswr well below 2:1, and yielding gains within 0.5 dB of those theoretically obtainable from the transistors used in the circuits. In fact, these designs were so successful that they ultimately formed the basis for a commercial product line for the 1296 MHz band. There is, however, no reason why full maximum available gain, and true 1:1 vswr, should not be achievable in microstripline amplifiers if simplistic design methods are abandoned in favor of a more complex, rigorous approach. In military and aerospace electronics, where the ultimate in performance is essential, such perfectionism has become a watchword. Small wonder, then, that a few discerning radio amateurs are attempting to achieve state-of-the-art performance in their microwave transistor designs.

This article is an exposition of state-of-the-art design techniques and is intended to be instructional, rather than constructional. By briefly reviewing the design approach applied in my previous articles, I shall attempt to identify the omissions which cause the resulting amplifier performance to depart from optimum. Following a brief discussion of transistor parameter characterization, I will show the mathematical procedures necessary to overcome the performance shortcomings of my previous designs. This is followed by a design example based on the MRF-901 transistor. The final section of the article outlines the minor performance differences between the rigorous design approach and more casual matching computations. For those readers who are interested, an appendix is provided which outlines some of the rules of vector arithmetic that are required for proper manipulation of semiconductor scattering parameters.

simple matching scheme

For any two-port device, optimum power transfer occurs when both input and output are terminated in their complex conjugate impedances. That is, for any port impedance consisting of a resistance and a series reactive component, the termination should appear as a like value of resistance in series with the *opposite* reactive component. For example, a port with an impedance of 35 + j100 ohms (35 ohms of resistance, 100 ohms of series *inductive* reactance), should be terminated in 35 − j100 ohms (35 ohms of resistance, 100 ohms of series *capacitive* reactance).

Recognizing the above, amplifier design consists of causing the ports (transistor input and output imped-

By H. Paul Shuch, WA6UAM, Microcomm, 14908 Sandy Lane, San Jose, California 95124

ances) to be terminated in their complex conjugates. The problem is to identify the actual transistor input and output impedances occurring at a given frequency, and under a given set of bias conditions.

It is convenient to assume that the impedances related to the transistor's input and output reflection coefficients (s_{11} and s_{22}) approximate the device's input and output impedances. This is the approach I used in my previous articles. The fallacy (and one source of minor errors) is the fact that s_{11}, the *input reflection coefficient,* directly relates to input impedance *only* when the output is terminated in a pure resistance of 50 ohms. Similarly, s_{22}, *output reflection coefficient,* is directly related to the output impedance *only* when the input is terminated in a pure resistance of 50 ohms. In other words, varying the impedance match at one port will affect the impedance seen at the other port.

To understand the reason for this interaction, it's important to realize that no transistor is strictly a unilateral device. Any time a signal is injected into the output of a transistor amplifier, some signal will be discernible at the input. The physics of semiconductor construction allow for a feedback path which, though it may appear minimal, nontheless allows output matching to have an impact on input impedance, and vice versa.

If the input and output reflection coefficients of a transistor were both zero, this feedback path would have no effect on transistor matching. A reflection coefficient of zero indicates a corresponding port impedance of 50 ohms, nonreactive. Since s_{11} is measured with the output terminated in 50 ohms, and s_{22} is measured with the input similarly loaded, this is the only case in which terminating one port will not disrupt matching to the other. Of course, if both the input and output impedances of a transistor were 50 ohms, pure resistive, matching to a 50-ohm source and load would be considerably simplified. Unfortunately, we are not blessed with such transistors. Hence, to properly determine input and output impedances, the device's *transfer coefficients* must be considered.

S-parameters

In addition to s_{11} and s_{22}, the reflection coefficients discussed previously, a microwave transistor is characterized by a *forward transfer coefficient,* s_{21} (mathematically related to gain), and a *reverse transfer coefficient,* s_{12} (which describes the internal feedback path). Together, these four scattering parameters fully characterize the operation of the device. From them can be calculated the transistor's stability factor (tendency to oscillate under various conditions of source and load termination), maximum available gain, maximum stable gain, and equivalent input and output impedances. The s-parameters can be further manipulated to determine the device's maximum linear power output capability[3] although such an analysis is beyond the scope of this article.

It should be remembered that each of the four s-parameters varies with frequency, as well as with varying conditions of bias current and operating potential. The term "scattering" is derived from the fact that the parameters describe a set of variables, based on traveling waves incident on a port and reflected (or scattered) from it, which are evaluated with a mathematical tool called a scattering matrix.[4]

It should be further pointed out that s-parameters are *vectors.* That is, they appear as points on a Smith chart or polar plot which can be defined by both *magnitude* and *angle.* For example, at a frequency of 1.3 GHz, with

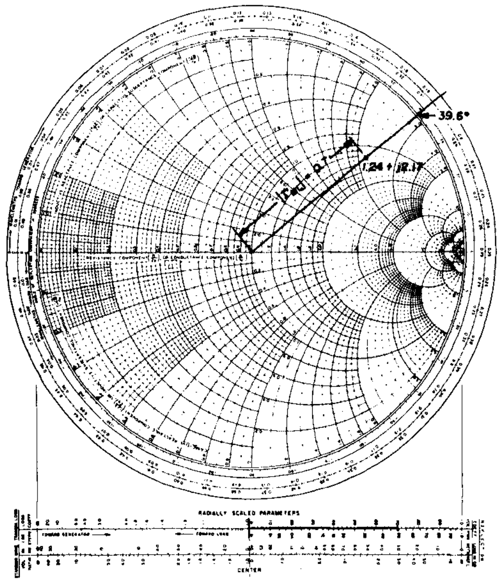

fig. 1. Using a Smith chart to plot the load impedance which exhibits the specified load reflection coefficient, 0.7 ∠ 39.6°. On this normalized Smith chart this yields 1.24 + j2.17. In a 50-ohm system the required load impedance is 62 + j108.5. To plot this point first locate the angle of the reflection coefficient on the peripheral scale and draw a line from 39.6° on this scale through the center of the chart. Referring to the radially scaled voltage reflection coefficient below the chart, measure the distance to 0.7, and transfer this length to the previously plotted line on the Smith chart. The crossover point marks the required complex load impedance.

a collector current of 10 mA and a collector-to-emitter potential of 10 volts, the common-emitter s-parameters for a Motorola MRF-901 microwave transistor are:

$$s_{11} = 0.47 \angle +161^\circ$$
$$s_{22} = 0.43 \angle -41^\circ$$
$$s_{12} = 0.08 \angle +64^\circ$$
$$s_{21} = 3.1 \angle +63^\circ$$

A complete discussion of the derivation and usefulness of the four s-parameters is available in an application note published by Hewlett-Packard.[5] Tabulations of s-parameters corresponding to various frequencies and bias conditions are available from the manufacturers of most microwave transistors.

gain and stability analysis

Before attempting to determine input and output impedances and design matching networks, it is desirable to approximate the gain capabilities of the transistor

under the chosen operating conditions, and to determine whether the resulting amplifier will be stable. Three parameters which aid in such analysis are Maximum Available Gain *(MAG)*, Maximum Stable Gain *(MSG)*, and Rollett's stability factor *(K)*. K indicates the amplifier's tendency to oscillate. If K is greater than 1, the amplifier will be stable under any combination of input and output impedances or phase angles. Such an amplifier is said to be *unconditionally* stable. Conservative

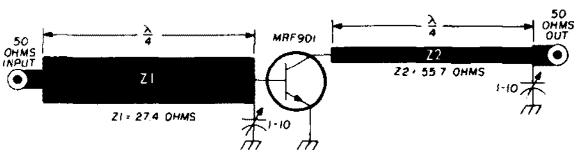

fig. 2. Basic circuit for a 1296-MHz amplifier which uses a Motorola MRF-901 transistor. Input and output matching is provided by microstriplines and 10 pF trimmer capacitors. This amplifier is unconditionally stable and gain is about 13 dB (bias networks are not shown).

design philosophy suggests that if K calculates to less than unity a different transistor or bias condition should be selected.

Maximum stable gain is, to quote WA6RDZ, ". . . the most important figure of merit. Transistors with high MSG are easy to match, easy to tune, and give high performance, trouble-free amplifiers."[6] Maximum available gain, also easily calculated, is a fairly accurate approximation of the gain you will observe in the actual circuit if it is carefully designed and built. If MAG is on the order of 2 or 3 dB less than MSG, the amplifier is likely to be both stable and reliable.

Of the above parameters, MSG is the most readily computed because it involves only the absolute values (magnitudes) of s_{21} and s_{12}:

$$MSG\,(dB) = 10 \log \frac{|s_{21}|}{|s_{12}|} \tag{1}$$

In order to perform the remaining calculations, the vector quantity *Δ and the scalar values B_2 are required:

$$\Delta = s_{11}\, s_{22} - s_{12}\, s_{21} \tag{2}$$

$$B_2 = 1 + |s_{22}|^2 - |s_{11}|^2 - |\Delta|^2 \tag{3}$$

It is now possible to calculate Rollett's stability factor, K

$$K = \frac{1 + |\Delta|^2 - |s_{11}|^2 - |s_{22}|^2}{2 \cdot |s_{21}| \cdot |s_{12}|} \tag{4}$$

If K proves greater than unity, go ahead and calculate maximum available gain:

$$MAG\,(dB) = MSG + 10 \log\, |K \pm \sqrt{K^2 - 1}| \tag{5}$$

where if B_2 is greater than zero (i.e., positive), the sign preceding $\sqrt{K^2 - 1}$ is negative, and if B_2 is less than zero (i.e., negative), the sign is positive. At this point in the circuit design, it is possible to determine whether the performance of the amplifier is acceptable for the intended application. If the amplifier proves only conditionally stable ($K \leq 1$), or if MAG is insufficient, select another transistor or bias point, and go through the calculations again with the new s-parameters.

*Rules for vector arithmetic are discussed in *appendix 1*.

output conjugate matching

Assuming that the gain and stability analysis indicate that the amplifier design is workable, the output circuit is designed to terminate the transistor in the complex conjugate of its *actual* output impedance. To determine the true output impedance requires a manipulation involving not only s_{22}, but also Δ and B_2 (eqs. 2 and 3) as well as s_{11}. To find the desired load reflection coefficient, first compute the intermediate vector quantity C_2:

$$C_2 = s_{22} - (\Delta \cdot s_{11}{}^*) \tag{6}$$

where the asterisk indicates that the complex conjugate of the immediately preceding vector is used (that is, same magnitude, angle has opposite sign).

The angle of the desired load reflection coefficient, $\Gamma_{ML\theta}$, is simply $C_{2\theta}{}^*$. The desired magnitude is found from:

$$|\Gamma_{ML}| = \frac{B_2 \pm \sqrt{B_2{}^2 - 4\,|C_2|^2}}{2\,|C_2|} \tag{7}$$

The sign preceding the radical sign is, once again, opposite to the sign on B_2. The desired load reflection coefficient may now be converted on a Smith chart into a complex impedance value, then matched to 50 ohms, as discussed in my previous articles.

input conjugate matching

Once the output load has been specified, the source reflection coefficient which will properly terminate the transistor's input is found from

$$\Gamma_{MS} = \left[s_{11} + \left(\frac{s_{12} \cdot s_{21} \cdot \Gamma_{ML}}{1 - (\Gamma_{ML} \cdot s_{22})} \right) \right]^* \tag{8}$$

where the asterisk indicates the complex conjugate (same magnitude, angle has opposite sign). This reflection coefficient may be plotted on a Smith chart to determine equivalent impedance, and the result transformed to 50 ohms.

To those readers who have Hewlett-Packard HP-45 engineering calculators, I highly recommend an article by Martin[7] which reduces formulas similar to the above to straightforward keystroke sequences. I have also published a HP-35 algorithm for series-to-parallel complex impedance conversion which may prove useful in designing matching networks.[8] Additionally, I recently derived a family of programs for the new HP-25 programmable calculator which greatly simplify all of the above calculations.*

*A complete set of HP-25 amplifier design programs is available from the author for $2.00 plus a stamped, self-addressed envelope.

design example

As noted previously, the common-emitter s-parameters for the Motorola MRF-901 transistor at 1.3 GHz, when biased at 10 volts and 10 mA, are as follows:

$$s_{11} = 0.47\angle{+161^\circ}$$
$$s_{22} = 0.43\angle{-41^\circ}$$
$$s_{12} = 0.08\angle{+64^\circ}$$
$$s_{21} = 3.10\angle{+63^\circ}$$

Using these parameters, the maximum stable gain, *MSG*, is calculated from **eq. 1**:

$$MSG\,(dB) = 10 \log \frac{s_{21}}{s_{12}}$$
$$= 10 \log (3.1/0.08) = 15.9\ dB$$

Before performing the remaining calculations, it's necessary to compute the vector quantity Δ (**eq. 2**) and the scalar quantity B_2 (**eq. 3**).

$$\Delta = (s_{11} \cdot s_{22}) - (s_{12} \cdot s_{21})$$
$$= [|s_{11}| \cdot |s_{22}| \angle(s_{11\theta} + s_{22\theta})] - [|s_{12}| \cdot |s_{21}| \angle(s_{21\theta} + s_{12\theta})]$$
$$= [|0.47| \cdot |0.43| \angle 161^\circ + (-41^\circ)] - [|0.08| \cdot |3.1| \angle(63^\circ + 64^\circ)]$$
$$= (0.202 \angle 120^\circ) - (0.25 \angle 127^\circ)$$

Converting to rectangular notation and subtracting the x and y components,

$$\Delta_x = 0.049 \qquad \Delta_y = -0.025$$

Returning to polar notation

$$\Delta_{tR} = \sqrt{\Delta_x^2 + \Delta_y^2} = \sqrt{0.00306} = 0.055$$
$$\Delta_{t\theta} = \arctan \Delta_y/\Delta_x = \arctan -0.500 = -26.56^\circ$$
$$\Delta = 0.055 \angle{-26.56^\circ}$$

The scalar quantity B_2 is calculated from the relationship

$$B_2 = 1 + |s_{22}|^2 - |s_{11}|^2 - |\Delta|^2$$
$$= 1 + (0.43)^2 - (0.47)^2 - (0.055)^2 = 0.96$$

With the vector quantity Δ and scalar quantity B_2 now known, it's possible to calculate Rollett's stability factor, *K*, from **eq. 4**.

$$K = \frac{1 + |\Delta|^2 - |s_{11}|^2 - |s_{22}|^2}{2\,|s_{21}| \cdot |s_{12}|}$$
$$= \frac{1 + (0.055)^2 - (0.47)^2 - (0.43)^2}{2(3.1)\,(0.08)} = 1.20$$

Since the stability factor is greater than 1, the amplifier will be unconditionally stable.

The maximum available gain, *MAG*, of the amplifier is calculated with **eq. 5**.

$$MAG\,(dB) = MSG\,(dB) + 10 \log |K \pm \sqrt{K^2 - 1}|$$

Since B_2 is greater than zero (i.e., positive) the sign of the radical is minus.

$$MAG\,(dB) = 15.9\ dB + 10 \log |1.20 - \sqrt{1.20^2 - 1}|$$
$$= 15.9\ dB + 10 \log 0.54$$
$$= 15.9 + (-2.7) = 13.2\ dB$$

Since *MAG* is approximately 3 dB lower than *MSG*, the amplifier can be expected to tune easily.

output matching. To terminate the transistor in the complex conjugate of its output impedance, first compute the intermediate vector quantity C_2 from **eq. 6**.

$$C_2 = s_{22} - (\Delta \cdot s_{11}^*)$$

remembering that the angle of s_{11}^* has a sign opposite to that of s_{11} (in this case, $s_{11}^* = 0.47 \angle{-161^\circ}$

$$C_2 = (0.43 \angle{-41^\circ}) - [(0.054 \angle{-25.61^\circ}) \cdot (0.47 \angle{-161^\circ})]$$
$$= (0.43 \angle{-41^\circ}) - (0.02 \angle 173.4^\circ)$$

Converting to rectangular notation and subtracting the x and y components,

$$C_{2x} = 0.344 \qquad C_{2y} = -0.284$$

Returning to polar notation

$$C_{2R} = \sqrt{C_{2x}^2 + C_{2y}^2} = \sqrt{0.199} = 0.45$$
$$C_{2\theta} = \arctan C_{2y}/C_{2x} = \arctan -0.826 = -39.6^\circ$$
$$C_2 = 0.45 \angle{-39.6^\circ}$$

The angle of the load reflection coefficient, $\Gamma_{ML\theta}$ is $C_{2\theta}^*$ where the asterisk indicates that the sign of the angle is changed. In this case, $C_{2\theta}^* = +39.6^\circ$. The magnitude of the load reflection coefficient is found from **eq. 7**:

$$|\Gamma_{ML}| = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2|C_2|}$$

where the sign ahead of the radical is opposite to the sign on B_2.

$$|\Gamma_{ML}| = \frac{0.96 - \sqrt{(0.96)^2 - 4(0.45)^2}}{2(0.45)}$$
$$= 0.70$$
$$\Gamma_{ML} = 0.70 \angle 39.6^\circ$$

This quantity may be plotted on a Smith chart to determine the desired load impedance as shown in fig. 1.* This yields

$$Z_L = 62.0 + j108.5$$

*The load impedance can also be calculated from the relationship

$$Z_L = Z_o \frac{1 + \Gamma_{ML}}{1 - \Gamma_{ML}}$$

where Z_L is the complex load impedance $(R \pm jX)$, $Z_o = 50 + j0$, and Γ_{ML} is the complex load reflection coefficient. Since all quantities are complex numbers, vector arithmetic is required.

Input matching. Now that the output load impedance has been specified, the source reflection coefficient which will properly terminate the input to the transistor can be calculated from **eq. 8**:

$$\Gamma_{MS} = \left[s_{11} + \left(\frac{s_{12} \cdot s_{21} \cdot \Gamma_{ML}}{1 - (\Gamma_{ML} \cdot s_{22})} \right) \right]^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{(0.08 \angle 64^\circ)(3.1 \angle 63^\circ)(0.7 \angle 39.6^\circ)}{1 - [(0.7 \angle 39.6^\circ)(0.43 \angle -41^\circ)]} \right)^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{(0.08 \cdot 3.1 \cdot 0.7) \angle 64^\circ + 63^\circ + 39.6^\circ}{1 - [(0.7 \cdot 0.43) \angle (39.6^\circ - 41^\circ)]} \right)^*$$

$$= (0.47 \angle 161^\circ) + \left(\frac{0.17 \angle 166.6^\circ}{0.70 \angle 0.60^\circ} \right)^*$$

$$= [(0.47 \angle 161^\circ) + (0.24 \angle 166^\circ)]^*$$

Converting to rectangular notation and adding the x and y components,

$$\Gamma_{MSx} = -0.68 \qquad I_{MSy} = 0.21$$

Returning to polar notation

$$\Gamma_{MSR} = \sqrt{\Gamma_{MSx}{}^2 + \Gamma_{MSy}{}^2} = \sqrt{0.51} = 0.71$$

$$\Gamma_{MS\theta} = \arctan \Gamma_{MSy}/\Gamma_{MSx} = \arctan -0.31$$

$$= 162.7^\circ$$

$$\Gamma_{MS} = 0.71 \angle 162.7^{\circ *} = 0.71 \angle -162.7^\circ$$

The source reflection coefficient for a complex conjugate input match may be plotted on a Smith chart to determine the corresponding source impedance. This yields the series complex impedance, $Z_s = 8.7 - j7.4$ ohms (parallel complex impedance, $Z_p = 15 \| -j17.6$ *ohms*).

Matching networks. An input conjugate match can be obtained by shunting the transistor base with a capacitive reactance equal to the desired parallel equivalent reactance value (-j17.6 ohms) and transforming the source impedance to the required parallel resistance value (15 ohms) through a quarter-wavelength transmission line. The required capacitance value is found from the familiar reactance equation

$$C = \frac{1}{2\pi f X_c}$$

At 1296 MHz:

$$C = \frac{1}{2\pi(1296 \cdot 10^6)\, 17.6} = 6.98\ pF$$

A 10 pF trimmer capacitor will assure a proper reactive termination.

Transformation of the resistive component (to a 50-ohm source in this case) is accomplished with a quarter-wavelength transmission line which has a characteristic impedance Z_o, equal to the geometric mean of the source impedance Z_f, and the parallel input resistance, R_p.

$$Z_o = \sqrt{R_p \cdot Z_f}$$

With a 50-ohm source and a parallel input resistance of 15 ohms

$$Z_o = \sqrt{15 \cdot 50} = 27.4\ ohms$$

At 1296 MHz this is easily provided by a microstrip transmission line 0.26 inches (6.5mm) wide and 1.16 inches (29.5mm) long on a 1/16-inch (1.5mm) double-clad, fiberglass printed-circuit board.

A similar quarter-wavelength transformer can be designed to match the resistive component, R_s, of the complex series output impedance (62 ohms) to a 50-ohm termination, Z_t. As before, the characteristic impedance of the transmission line is given by

$$Z_o = \sqrt{R_s \cdot Z_t} = \sqrt{62 \cdot 50} = 55.7\ ohms$$

At 1296 MHz this is provided by a microstrip transmission line 0.09 inches (2.3mm) wide and 1.21 inches (30.7mm) long on 1/16-inch (1.5mm) double-clad, fiberglass printed-circuit board.

The required inductive reactance in series with the collector (+j108.5 ohms) is provided by shunting a capacitive reactance across the output end of the quarter-wavelength transformer. The required capacitive reactance is given by

$$X_c = \frac{Z_o{}^2}{X_s} = \frac{55.7^2}{108.5} = 28.6\ ohms$$

At 1296 MHz:

$$C = \frac{1}{2\pi(1296 \cdot 10^6)\, 28.6} = 4.3\ pF$$

Again, a 10 pF trimmer capacitor will suffice. The circuit in fig. 2 shows the complete matching layout.

performance comparison

A simplified amplifier design, in which source and load impedances appear as the complex conjugate of the impedances related to s_{11} and s_{22}, yields the circuit shown in fig. 3. This circuit was derived by matching to the following assumed shunt-equivalent impedances:

Parallel: Z_{in} *(derived from* s_{11}*)*

$$= 21 \| j56.5\ ohms$$

Series: Z_{out} *(derived from* s_{22}*)*

$$= 75 - j52.5\ ohms$$

A more rigorous analysis shows the actual device impedances to be:

Parallel: $Z_{in}\ (actual) = 15 \ \| +j17.6\ ohms$

Series: $Z_{out}\ (actual = 62.0 - j108.5\ ohms$

Note that the reactive components of the shunt input impedance and the series output impedance differ significantly. Thus some degree of mismatch can be anticipa-

ted if the circuit of **fig. 3** is built as shown. Since the actual device impedances are now known, this mismatch can be accurately predicted.

As it happens, only the resistive component of the transistor's input or output complex impedance sees a mismatch. This is because the tuning range of the trimmer capacitors in **fig. 3** is sufficiently wide to properly terminate the reactive components. The input and output mismatches are determined by transforming the actual resistive components through the existing quarter-wave length transformers and comparing the resulting impedance to 50 ohms. Referring to **fig. 3**,

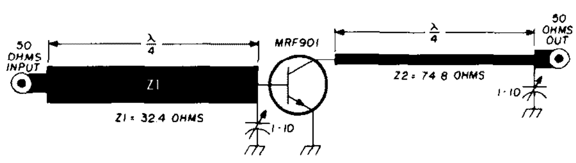

fig. 3. 1296-MHz amplifier which was designed with a simplified method. Input and output matching is provided by microstriplines and 10 pF trimmer capacitors. This amplifier is unconditionally stable and has about 0.5 dB less gain than the circuit shown in fig. 2 (bias networks are not shown).

$$Z_{in}\ (amplifier) = \frac{Z_1^{\,2}}{R_{p\ in}} = \frac{32.4^2}{15.0} = 70.0\ ohms$$

$$Z_{out}\ (amplifier) = \frac{Z_1^{\,2}}{R_{s\ out}} = \frac{74.8^2}{62.0} = 90.2\ ohms$$

Thus, the input vswr is 1.4:1 and the output vswr is 1.8:1, calculated values which correlate quite closely with those values observed in the actual amplifiers.

These input and output mismatches will result in somewhat lower stage gain than available from a properly terminated device Actual stage gain is found from

$$A_p\ (dB) = MAG + G_1 + G_2$$

where G_1 and G_2 are both negative and represent the mismatch losses at the input and output, respectively. Since G_1 (for a 1.4:1 vswr) is about -0.1 dB, and G_2 (for a 1.8:1 vswr) is about -0.4 dB,

$$A_p = 13.2 + (-0.1) + (-0.4)$$
$$= 12.7\ dB$$

This closely represents the measured gain of the amplifier shown in **fig. 3**.

summary

A method has been outlined for using device s-parameters to analyze the gain and stability of a microwave amplifier, and to determine appropriate source and load impedances for a complex conjugate match. It has been shown that designing around the reflection coefficients of a particular transistor (while ignoring the transfer coefficients) resulted in input and output mismatches of 1.4:1 and 1.8:1, respectively, while degrading overall amplifier gain by approximately 0.5 dB. This is a modest penalty for enjoying the convenience of a simplistic design approach. Whether the additional performance available from the more rigorous design method is justified depends largely upon the goals of the designer, and the intended application of the amplifier.

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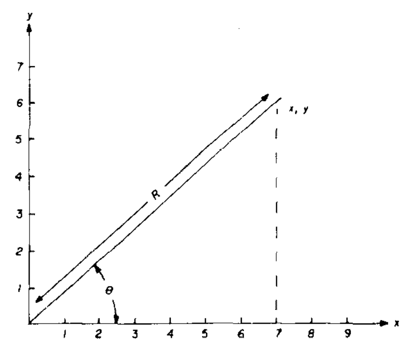
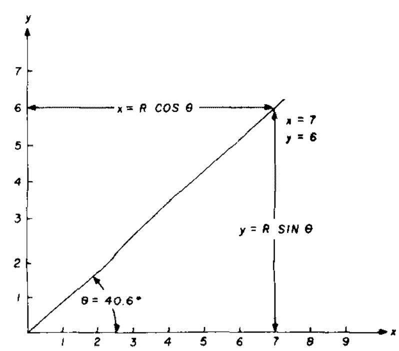
appendix 1

vector arithmetic

In the application of s-parameter design equations, it's necessary to perform numerous computations involving vector quantities. Vectors may be expressed either in conventional polar notation (magnitude R and associated angle θ) or may be resolved into their rectangular



components (x and y displacement on Cartesian coordinates) as shown below. Since the vector is part of a right triangle, manipulation between the two forms of notation involves the application of trigonometric functions. These manipulations may be accomplished on a slide rule, manually with the aid of trig tables, or on a hand-held digital calculator.



Readers who have advanced scientific calculators which include polar-rectangular conversion and summation keys will find the process considerably simplified. The following review is for the benefit of those not so fortunate.

1. Resolving vectors. Vector arithmetic often requires that the x and y components of the vector be known. Any vector, V, described by magnitude, R, and angle θ, can be resolved into its x and y components with the following formulas

$$x = R \cos \theta$$
$$y = R \sin \theta$$

Example: What are the x and y components of the vector $9.22 \angle 40.6^\circ$ ($R = 9.22$, $\theta = 40.6^\circ$)?

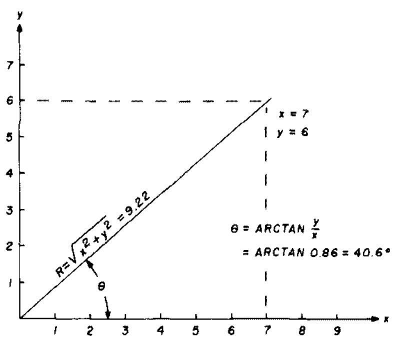
$$x = R \cos \theta = 9.22 \cos 40.6^\circ$$
$$= 9.22 \cdot 0.76 = 7.00$$
$$y = R \sin \theta = 9.22 \sin 40.6^\circ$$
$$= 9.22 \cdot 0.65 = 6.00$$

2. Constructing vectors. The results of vector addition and subtraction (reviewed later) generally appear as x and y components of the resultant vector. These coordinates can be converted to a polar vector of magnitude, R, and angle, θ, with measurements on a graphical plot, or by using a trigonometric solution. Since the vector is represented by the hypotenuse of a right triangle formed by dimensions x and y, the Pythagorean theorem may be used to find the magnitude R:

$$R = \sqrt{x^2 + y^2}$$

Trigonometry is used to calculate the angle θ

$$\theta = \arctan y/x$$



Example: What is the magnitude, R, and angle θ, of the vector described by the Cartesian coordinates, $x = 7$, $y = 6$?

$$R = \sqrt{x^2 + y^2} = \sqrt{49 + 36}$$
$$= \sqrt{85} = 9.22$$
$$\theta = \arctan y/x = \arctan 6/7$$
$$= \arctan 0.86 = 40.6^\circ$$

3. Vector addition. Any two vectors V_1 and V_2, which have been resolved into x and y components V_{1x}, V_{1y}, V_{2x}, and V_{2y}, may be added by summing the respective x and y components. The summed components may then be constructed into a resultant vector V_t, of magnitude V_{tR} and angle $V_{t\theta}$.

Example: What is the sum of vectors V_1 and V_2 when

$$V_1 = 1.5 \angle +40^\circ$$
$$V_2 = 2.0 \angle -60^\circ$$
$$V_{1x} = V_{1R} \cos V_{1\theta} = 1.5 \cos 40^\circ = 1.15$$
$$V_{1y} = V_{1R} \sin V_{1\theta} = 1.5 \sin 40^\circ = 0.96$$
$$V_{2x} = V_{2R} \cos V_{2\theta} = 2.0 \cos -60^\circ = 1.00$$
$$V_{2y} = V_{2R} \sin V_{2\theta} = 2.0 \sin -60^\circ = -1.73$$
$$\Sigma_x = V_{1x} + V_{2x} = 1.15 + 1.00 = 2.15$$
$$\Sigma_y = V_{1y} + V_{2y} = 0.96 + (-1.73) = -0.77$$
$$V_{tR} = \sqrt{\Sigma_x{}^2 + \Sigma_y{}^2} = \sqrt{2.15^2 + (-0.77)^2}$$
$$= \sqrt{5.22} = 2.28$$
$$V_{t\theta} = \arctan (\Sigma_y/\Sigma x) = \arctan (-0.77/2.15)$$
$$= -19.70^\circ$$
$$V_{t\theta} = V_1 + V_2 = 2.28 \angle -19.70^\circ$$

4. Vector subtraction. Any two vectors V_1 and V_2, which have been resolved into x and y components V_{1x}, V_{1y}, V_{2x}, and V_{2y}, may be subtracted one from the other by subtracting their respective x and y components. The results of such subtraction comprise the x and y components of the resulting vector V_t, which may be constructed to yield magnitude V_{tR} and angle $V_{t\theta}$.

Example: What is the difference when vector V_2 is subtracted from vector V_1?

$$V_1 = 0.8 \angle -40^\circ$$
$$V_2 = 0.4 \angle +120^\circ$$
$$V_{1x} = V_{1R} \cos V_{1\theta} = 0.8 \cos -40^\circ = +0.61$$
$$V_{1y} = V_{1R} \sin V_{1\theta} = 0.8 \sin -40^\circ = -0.51$$
$$V_{2x} = V_{2R} \cos V_{2\theta} = 0.4 \cos 120^\circ = -0.20$$
$$V_{2y} = V_{2R} \sin V_{2\theta} = 0.4 \sin 120^\circ = +0.35$$
$$(\Sigma_{-x}) = V_{1x} - V_{2x} = +0.61 - (-0.20) = +0.81$$
$$(\Sigma_{-y}) = V_{1y} - V_{2y} = -0.51 - (+0.35) = -0.86$$
$$V_{tR} = \sqrt{(\Sigma_{-x})^2 + (\Sigma_{-y})^2} = \sqrt{(0.81)^2 + (-0.86)^2}$$
$$= \sqrt{1.40} = 1.18$$
$$V_{t\theta} = \arctan (\Sigma_{-y})/(\Sigma_{-x}) = \arctan (-0.86/+0.81)$$
$$= -46.7^\circ$$
$$V_t = V_1 - V_2 = 1.18 \angle -46.7^\circ$$

5. Vector multiplication. For any two vectors, V_1 and V_2, each described by a magnitude, R, and an angle, θ, the vector product is found by multiplying the the magnitudes and adding the angles:

$$V_{tR} = V_{1R} \times V_{2R}$$
$$V_{t\theta} = V_{1\theta} + V_{2\theta}$$

Example: What is the vector product of V_1 and V_2 when

$$V_1 = 0.8 \angle 45^\circ$$
$$V_2 = 0.65 \angle -118^\circ$$
$$V_{tR} = V_{1R} \times V_{2R} = 0.8 \times 0.65 = 0.52$$
$$V_{t\theta} = V_{1\theta} + V_{2\theta} = +45^\circ + (-118^\circ) = -73^\circ$$
$$V_t = V_1 \cdot V_2 = 0.52 \angle -73^\circ$$

6. Vector division. For any two vectors, V_1 and V_2, each described by a magnitude, R, and an angle, θ, the vector quotient is found by dividing the magnitudes and subtracting the angles:

$$V_{tR} = V_{1R} \div V_{2R}$$
$$V_{t\theta} = V_{1\theta} - V_{2\theta}$$

Example: What is the vector quotient when V_1 is divided by vector V_2?

$$V_1 = 0.96 \angle +64^\circ$$
$$V_2 = 0.42 \angle -102^\circ$$
$$V_{tR} = V_{1R} \div V_{2R} = 0.96 \div 0.42 = 2.29$$
$$V_{t\theta} = V_{1\theta} - V_{2\theta} = +64^\circ - (-102^\circ) = +166^\circ$$
$$V_t = 2.29 \angle +166^\circ$$

7. Maximum angle. Whenever a vector manipulation yields an expression whose angle exceeds ±180°, subtract the absolute value of the angle from 360°, and assign to the resulting angle a sign opposite to that of the original angle.

Examples: $-196^\circ = +360 - |-196|$

$$= +360 - 196 = +164^\circ$$

$$265^\circ = -360 - |+265|$$

$$= -360 - 265 = -95^\circ$$

8. Compound expressions. In expressions involving both vector and scalar quantities, treat the scalar quantity as though it were a vector of angle 0°.

Example: In the expression for the source reflection coefficient which will properly terminate the transistor's input (eq. 8), the product of the load reflection coefficient, Γ_{ML}, and s_{22} are subtracted from one. If $(\Gamma_{ML} \cdot s_{22}) = 0.65 \angle -118^\circ$, what is the value of the expression, $1 - (\Gamma_{ML} \cdot s_{22})$?

$$V_1 = 1 \angle 0^\circ$$

$$V_2 = 0.65 \angle -118^\circ$$

$$V_{1x} = V_{1R} \cos V_{1\theta} = 1 \cos 0^\circ = 1.00$$

$$V_{1y} = V_{1R} \sin V_{1\theta} = 1 \sin 0^\circ = 0$$

$$V_{2x} = V_{2R} \cos V_{2\theta} = 0.65 \cos -118^\circ = -0.31$$

$$V_{2y} = V_{2R} \sin V_{2\theta} = 0.65 \sin -118^\circ = -0.57$$

$$\Sigma_{-x} = V_{1x} - V_{2x} = 1.00 - (-0.31) = 1.31$$

$$\Sigma_{-y} = V_{1y} - V_{2y} = 0 - (-0.57) = 0.57$$

$$V_{1R} = \sqrt{(\Sigma_{-x})^2 + (\Sigma_{-y})^2} = \sqrt{1.31^2 + 0.57^2}$$

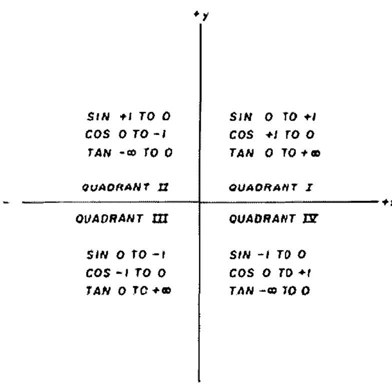
$$= \sqrt{2.04} = 1.43$$

$$V_{1\theta} = \arctan (\Sigma_{-y}/\Sigma_{-x}) = \arctan 0.57/1.31$$

$$= 23.51^\circ$$

$$1 - (\Gamma_{ML} \cdot s_{22}) = 1.43 \angle 23.51^\circ$$

9. Angular functions. Since most trigonometry tables show only the functions to +90°, when working with vectors which may fall in any of the four quadrants below (0 through 360 degrees) this can lead to ambiguities in specifying the angle θ of a resultant vector. Note that the



tangent function varies from zero to +∞, from zero to 90 degrees, from -∞ to zero in the second quadrant (90 to 180 degrees), from zero to +∞ in the third quadrant (180 to 270 degrees), and from -∞ to zero in the fourth quadrant (270 through 360 degrees). The sine and cosine functions are also ambiguous, as shown, but this doesn't create a problem in vector arithmetic.

In the expression for the angle of the source reflection coefficient, $\Gamma_{MS\theta}$, in the design example $\Gamma_{MSx} = -0.68$ and $\Gamma_{MSy} = 0.21$. Therefore,

$$\Gamma_{MS\theta} = \arctan -0.68/0.21 = \arctan 0.31$$

$$= -17.3^\circ$$

This is in the fourth quadrant whereas the x and y values place the vector in the second quadrant. Therefore, the correct value for the angle is 180° + (-17.3°) = 162.7°. The same sort of ambiguity exists for the first and third quadrants, and can only be resolved by inspection.

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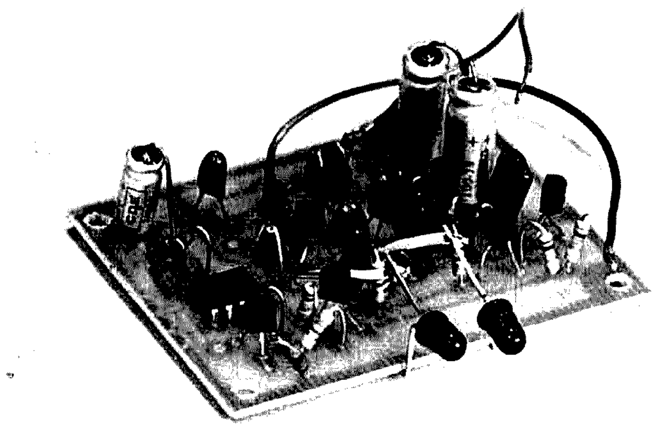


two-channel scanner

for repeater monitoring

Simple modifications to WA2GCF's popular channel scanner for use with the Icom IC-2F and other vhf-fm transceivers

The two-channel scanner described in this article was never really designed — it evolved. I wanted a simple scanner to monitor two repeaters, using a small solid-state transceiver. The simplest circuit I found was one described by WA2GCF in an article in *ham radio*.[1] The unit proved to have all the advantages the author claimed; it was inexpensive, small, easy to assemble, and operates directly from the 13-volt transceiver power supply or an auto battery. Only minor modifications were necessary to monitor the two club repeaters with my Icom IC-2F.

The first problem was how to use diode switching with the IC-2F receiver oscillator. WA2GCF described circuitry for switching crystals in oscillators requiring +5 volts to turn on, and oscillators requiring a ground for turn-on. Both are designed for circuits where one side of the crystal is grounded. The IC-2F, however, like many other Japanese receivers, has a receiver oscillator in which the crystal is connected between the collector and base of the oscillator transistor (**fig. 1A**). Neither side of the crystal can be grounded. At first it appeared it would be necessary to redesign the oscillator circuit, but the switching circuit modification shown in **fig. 1B**, which required rewiring only the switch, proved useable.

squelch recognition

Scanning action is stopped when a carrier opens the receiver squelch. The original scanner could be stopped by a squelch circuit that goes to ground when an incoming signal is received, or with alternate wiring by a squelch circuit that is at ground potential when the squelch is closed and goes high when it is opened. Unfortunately, the IC-2F does neither. Its squelch circuit is high when the squelch is closed, and goes to a lower voltage when it opens, but not all the way to ground. The scanner stopped for strong signals but not for ones that were weak but still quite readable.

The original circuit used to stop the scanner multivibrator is modified by changing Q1 to a pnp transistor with its emitter connected to the positive supply voltage, and adding an npn transistor, Q1A, to control it as

By Pat Shreve, W8GRG, 2842 Winthrop Road, Shaker Heights, Ohio 44120

shown in fig. 2. The base of Q1A is biased so that it will be turned off when the squelch circuit connected to its emitter is high; it is turned on by a voltage less than the bias but still above ground potential, which occurs when many club members want to do. If the repeater carrier does not drop between users' transmissions, the scanner locks on that repeater and will stay there for the entire conversation. Without a periodic search-back feature

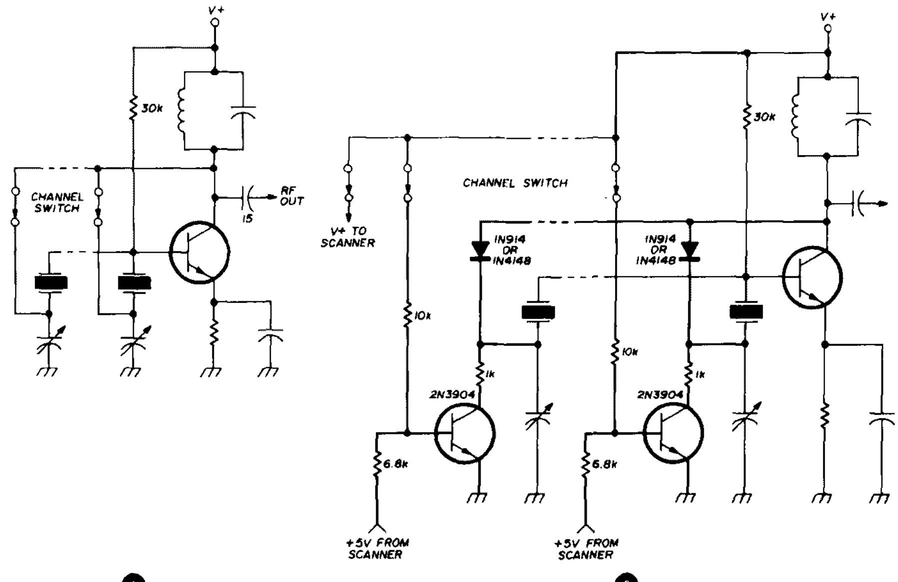


fig. 1. IC-2F receiver oscillator circuit. (A). As originally wired. (B). Modified for diode switching. Heavy lines snow wiring changes.

the squelch opens. When Q1A is off, Q1 and Q2 are also off, and the multivibrator can oscillate. When the three transistors turn on, the multivibrator is locked and scanning action stops.

The scanner still had one disadvantage for monitoring our two repeaters, which all our control operators and there is no way for the listener to know what may be happening on the other repeater.

To prevent lock-up on any channel for long periods a lock release, rather than a priority channel circuit, is needed. This can be done by adding a timer to the squelch recognition circuit described above. Instead of

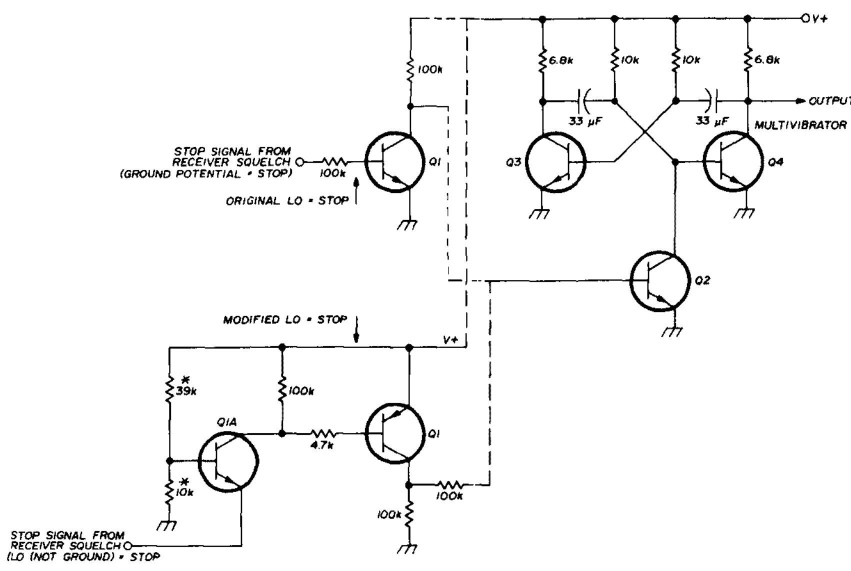


fig. 2. Multivibrator circuit modification to stop scanning action with low but above-ground stop signal.

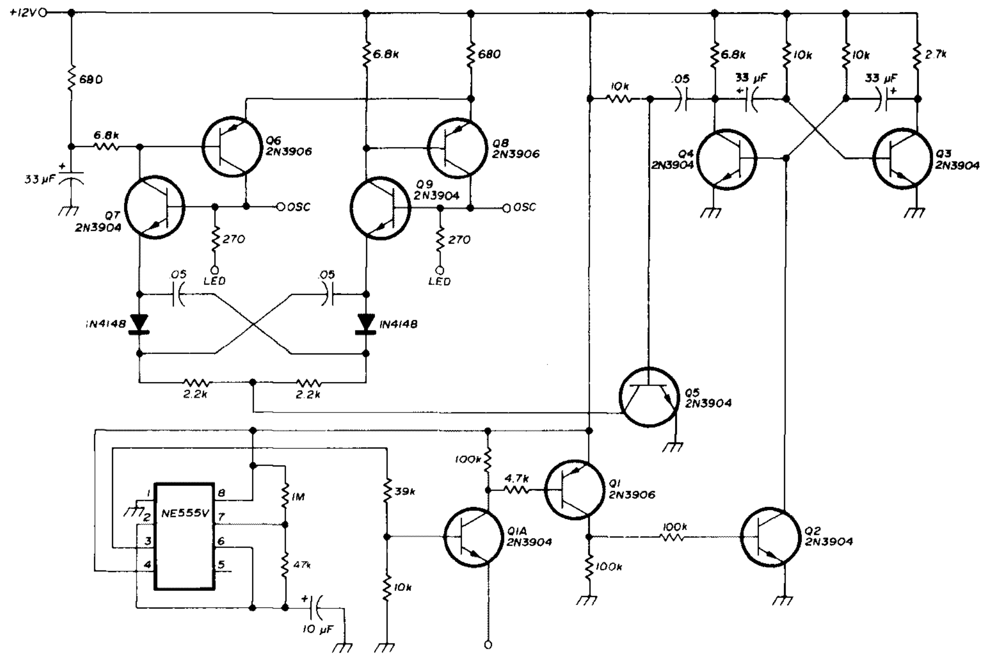


fig. 3. Two-channel scanner with search-back feature.

providing bias for Q1A from the V+, it is supplied from the output of an NE555 timer IC which is programmed to be high for about 15 seconds and low for a fraction of a second. Each time the IC cycles the scanner will move to the next frequency. With a two-channel scanner both frequencies are checked at least every 15 seconds. Even with four channels none is unguarded for more than a minute.

control switching

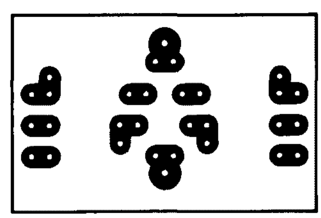
The scanner does not switch transmitter crystals, so manual switching to a selected frequency is necessary when the operator wants to talk. My IC-2F is a six-channel transceiver, with a six-position, double-pole, channel selector switch. As originally wired in **fig. 1A**, the common terminal on the receive side was connected to the collector of the oscillator. Rewired as in **fig. 1B**, this half of the switch controls V+ to the crystal switching circuits of four crystals. The other two positions, in which there are no crystals, control V+ to the scanner and the synthesizer, so that the scanner is disabled when the switch is turned to any transmitting position.

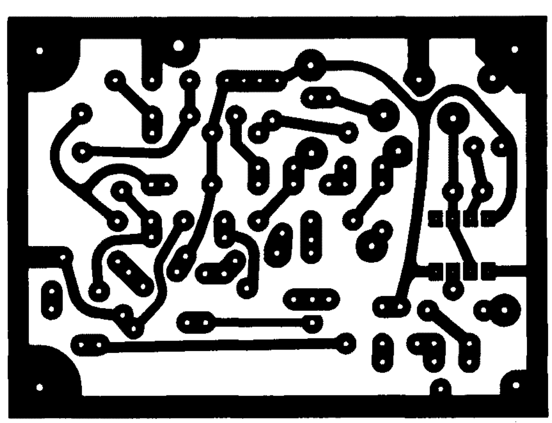
construction

The scanner modifications shown here can be easily applied to the scanner described in WA2GCF's article. For those who want only two channels, the complete circuit is reproduced in **fig. 3**. Full-size circuit board patterns for the scanner and the crystal-switching circuit are shown in **fig. 4**.* Component placement is shown in **figs. 5** and **6**. All of the components except the circuit boards can be purchased from *ham radio* advertisers for less than $5, or obtained locally. Tolerances are not critical, and almost any switching diodes and general-purpose silicon transistors can be substituted for those shown without affecting operation.

The 680-ohm resistor and 33-µF capacitor in the bias supply of Q6 unbalance the circuit sufficiently to insure

fig. 4. Full-size printed-circuit layouts for the switching circuit (top) and the scanner (below).





*A set of drilled, glass-epoxy, printed-circuit boards is available for $5, postpaid, from D.L. McClaren, 19721 Maplewood Ave., Cleveland, Ohio 44135

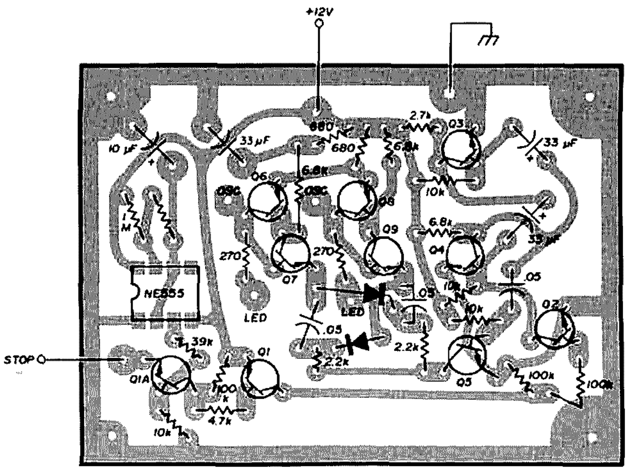


fig. 5. Printed-circuit component layout for the scanner viewed from the component side of board. Circles indicate external connections.

the start of scanning action. With the scan rate provided by the multivibrator circuit components shown in fig. 3, and the NE555 timing values, the scanner should stop every time it senses an active frequency.

adjustment

The scan rate can be varied by changing the values of the multivibrator capacitors. The length of time the scanner will stay locked on a single frequency can be

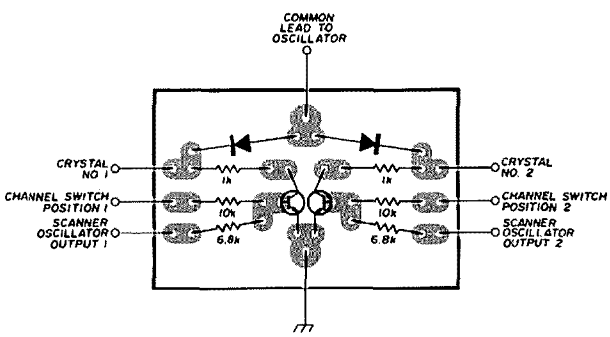


fig. 6. Printed-circuit component layout for oscillator switching circuit shown in fig. 1B.

changed by replacing the 1-megohm NE555 charging resistor with a different value. If these timing components are changed, it may be necessary to change the value of the 47k discharge resistor, also, to limit the multivibrator to only one cycle each time the NE555 output goes low. Excessive low time will cause the scanner to miss a signal on the next channel.

reference

1. Jerry Vogt, WA2GCF, "Improved Channel Scanner for VHF FM," *ham radio*, November, 1974, page 26.

ham radio

how to improve receiver performance of vacuum-tube vhf-fm equipment

Simple circuit modification for replacing 6AK5 receiver rf amplifier stages with a low noise, dual-gate, mosfet

Vhf fm operation is becoming increasingly popular with radio amateurs, with tens of thousands of fm transmitters and receivers now in use, and more being put into service daily. Not an insignificant number of these rigs are converted from commercial service – equipment that was originally manufactured by firms such as Motorola, RCA, and GE. However, most of the older receivers use vacuum-tube front ends so they don't provide the sensitivity that is possible with modern solid-state devices. This article describes a simple, proven mosfet circuit that can be easily substituted for those noisy tube-type front ends.

The most popular receiver rf amplifier used by communications equipment designers back in the late 1950s and early 1960s was based on the 6AK5 pentode – a workhorse used heavily during the war in vhf radar and communications systems. Typical receiver sensitivities provided by this tube are on the order of 1.5 to 3.0 microvolts for 20 dB quieting. Considering the state of the art for those days, that was not shabby performance.

In recent years the communications designer has been provided with a proliferation of very reasonably priced, high performance, dual-gate mosfet devices from several different manufacturers. One example is the 3N204 by Texas Instruments which has optimum spot noise figures of 2 dB at 200 MHz, 3 dB at 400 MHz, degrading to 7 dB at 900 MHz, all for $1.25 in small quantities. Anyone who has worked with a 417A/5842 vacuum-tube converter from the 1950s can certainly appreciate how rapidly technology has marched forward during the past two decades.

By Hank Meyer, W6GGV, 29330 Whitley Collins Drive, Rancho Palos Verdes, California 90274

It's a relatively simple task to replace the 6AK5 rf amplifier in your present rig with a 3N204 or other dual-gate device. Fig. 1 shows the circuit of a typical 6AK5 rf amplifier stage found in many commercial rigs, while fig. 2 shows the base diagram of the 6AK5. The following steps describe the circuit modification, which can be accomplished in less than an hour.

1. Remove the center pin from the 6AK5 tube socket by bending it over and breaking it with a pair of pliers. Cut any wires which are soldered to the center grounding pin.

2. Install a 33k, 1/2-watt resistor from pin 6 to ground. If there is already another resistor from pin 6 to ground,

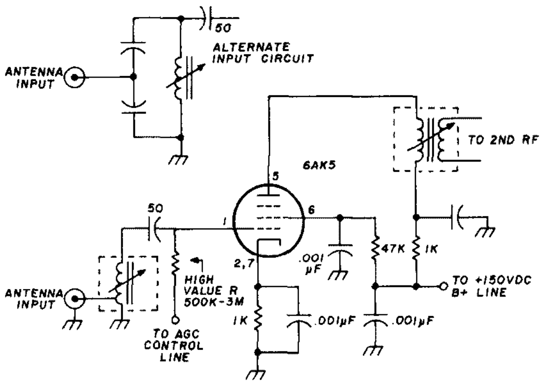


fig. 1. Typical 6AK5 receiver rf amplifier stage found in many older commercial vhf-fm rigs. Performance of this circuit is typically 1.5 to 3.0 microvolts for 20 dB quieting. Replacing the 6AK5 with a vhf dual-gate mosfet, as described here, increases sensitivity to 0.25 to 0.30 microvolt for 20 dB quieting.

remove it. Make sure that there is a 0.001 µF disc bypass capacitor between pin 6 and ground.

3. Remove the original resistor from the B+ line to pin 6 and replace it with a 120k, 1/2-watt resistor.

4. Remove the voltage-dropping resistor(s) from the B+ line to the B+ point on the 6AK5 rf amplifier plate output coil and replace it with a 300-ohm, 1/2-watt resistor.

5. Break the B+ line from the high-voltage feedpoint to the 6AK5. Now install a 3-lug terminal strip (two insulated lugs, one grounded lug) by soldering the ground lug to a convenient point on the chassis. Install a 6.8k, 2-watt resistor between the two insulated tie points and connect one end of the resistor to the previously removed B+ line. Connect an 18-volt zener diode (400 mW to 1 watt rating) from the other end of the 6.8k resistor to ground. From this same point connect a wire to the former B+ feedpoint for the 6AK5.

6. Insert the 3N204 mosfet into the center hole of the 6AK5 tube socket.

7. Using the 3N204 basing diagram in fig. 3, make the following connections to the 6AK5 tube socket:

Drain to pin 5
Source to ground (pin 3 or 4)
Gate 1 to pin 1
Gate 2 to pin 6

8. Disconnect the antenna input lead and reconnect it to the top of the rf input coil as shown in fig. 3.

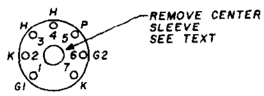


fig. 2. Base diagram (bottom view) of the 6AK5 tube socket.

9. Using a signal generator or on-the-air signal, repeak the tuned input and output circuits for maximum output.

10. Remove the 6-volt filament wiring from the socket, but before you do, check to see if two or more tubes are wired in series. If they are, install a 36-ohm, 2-watt resistor in series with the filament line to compensate for the current drain of the 6AK5 filament. This completes the conversion.

I have converted several 6AK5 rf amplifier stages to dual-gate mosfets using this simple procedure, and all have provided outstanding results. Sensitivity measurements using a Hewlett-Packard 608D signal generator indicate a sensitivity of about 0.25 to 0.30 microvolts for 20 dB quieting — a marked improvement in performance.

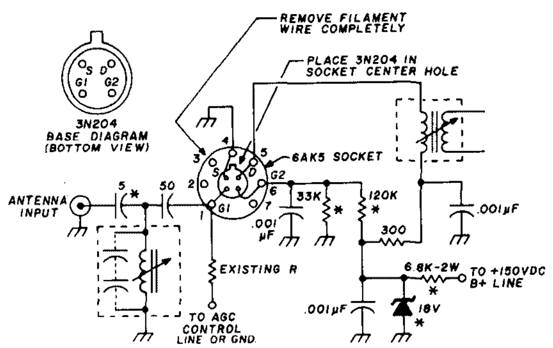


fig. 3. Modified receiver rf amplifier using a 3N204 dual-gate mosfet. Components marked with an asterisk are new.

In addition to the 3N204 I have tried several other, similar dual-gate devices, including the 3N201 and 3N200, all with good success. All of these devices are in the same price class, although the 3N204 has a bit better performance. If you're using an older tube-type commercial rig, this simple modification can significantly extend your receiving range and operating pleasure.

ham radio

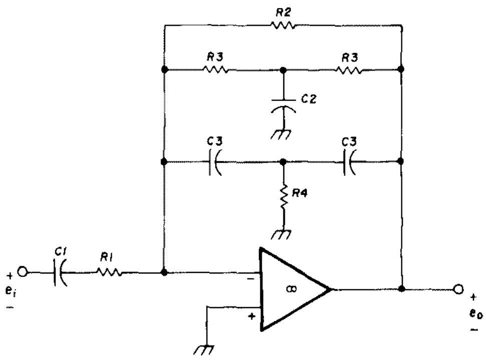
RC active filters using op amps

An overview of three of the most popular filter circuits,

A narrow bandpass filter of small size and low power consumption is often needed for increased receiver selectivity or other communications applications, such as RTTY. Active filters are also presently found in IC form for applications in modems, function generators, and level detectors.[1] This article presents information that has been omitted from the many articles on the subject in the amateur literature. Three of the most popular RC active filters are described together with their design equations and response characteristics. Also included is a discussion of breadboard testing and results of filter performance both singly and in cascaded form.

choice of circuit

The parallel-T network is a popular narrowband filter, which is used in the feedback loop of an op amp in various ways. In this case the network was inserted between the inverting input and op-amp output: In this



arrangement R1, C1 form part of the network, allowing a symmetrical response about center frequency. This added input network increases the lower-frequency skirt selectivity, which otherwise would not fall off as fast as the upper-frequency skirt characteristic. The frequency-response equation including phase relationships is:

$$\frac{e_o}{e_i}(\omega) = -\left\{\frac{1}{R1 + \frac{1}{j\omega C1}}\right\}\left\{\frac{R2[2 + j\omega C2R3]}{\left[1 - \omega^2\left(\frac{C2R3}{2}\right)^2\right]\frac{R2}{R3} + \left[2 + j\omega C2R3\right]}\right\}$$

The response equation remains in this form since it becomes rather cumbersome to deal with if expanded further. The design equations for the filter are:

$$f_o = \frac{1}{\pi C2R3}\sqrt{1 + \frac{2R3}{R2}} \approx \frac{1}{\pi C2R3} \tag{2}$$

$$C1 = \frac{\alpha}{2\pi f_o R1} \tag{3}$$

where $\alpha = \sqrt{2}$

$$R2 = R1 = \frac{1}{2R3(\pi f_o C2)^2} \tag{4}$$

for $\frac{e_o}{e_i} = 1$

$$R3 = 2R4 \tag{5}$$

Typical values chosen for this filter were:

R1 39,000 ohms	R4 1727 ohms	C3 0.05 μF
R2 39,000 ohms	C1 0.006 μF	f_o 1000 Hz
R3 3454 ohms	C2 0.1 μF	3-dB BW,
Q 14 (four stages); shape factor 13		(four Stages 71 Hz)

The Q can be made very high if desired, but the skirt

By Fred M. Griffee, W4IYB, 8809 Stark Road, Annandale, Virginia 22003

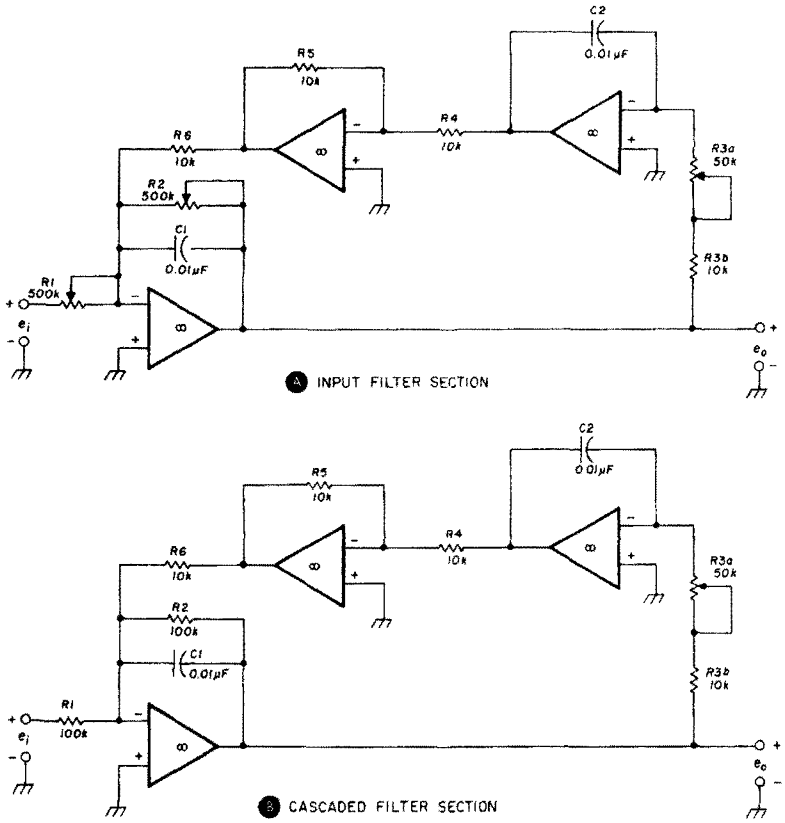
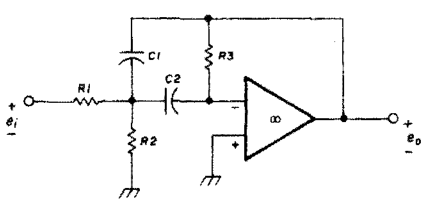


fig. 1. Input bandpass filter section, A, and cascaded sections, B, used for analysis. The rationale for component values is discussed in the text.

selectivity or bandpass-filter shape factor with only one stage remains poor. With three or four stages, the skirt selectivity and shape factor become fairly good for a narrow-bandpass audio filter; however, the component tolerance sensitivity is high.

Another circuit employing feedback in a slightly different manner (the same approach used by the popular MFJ-CWF2 filter[2]) to acquire the bandpass characteristic is given below (for one stage):



The response and design equations can be derived in the usual manner as in the previous case using flow-graph analysis. They are given below for reference:

$$\frac{e_o}{e_i}(\omega) = -\frac{j\omega[1/R1C1]}{\left[\frac{R1+R2}{R1R2R3C1C2} - \omega^2\right] + j\omega\left[\frac{C1/C2+1}{R3C1}\right]} \quad \textbf{(6)}$$

$$f_o = \frac{1}{2\pi\sqrt{\frac{R1+R2}{R1R2R3C1C2}}} \quad \textbf{(7)}$$

$$\frac{e_o}{e_i}(\omega_o) = -\frac{R3}{2R1} \quad \textbf{(8)}$$

$R1 = ½f_oC$, where $C = C1 = C2$

$R3 = \alpha/f_oC$

$R2 = 1(2\pi)^2 f_oC\alpha$

$Q \approx \pi\alpha$, letting $\alpha = \sqrt{2}$

Again, this filter will not have good skirt selectivity and cascaded stages must be used to obtain a good shape factor. Otherwise, regardless of Q, signals will still be heard with respect to sideband frequencies far removed from the center frequency, f_o. Values chosen from the design equations for this filter are given below for four stages:

R1 975,000 ohms	C1 0.001 µF	f_o 1000 Hz
R2 13,165 ohms	C2 0.001 µF	3-dB BW 71 Hz
R3 1.95 megohms		Q = 14
Shape factor 13		

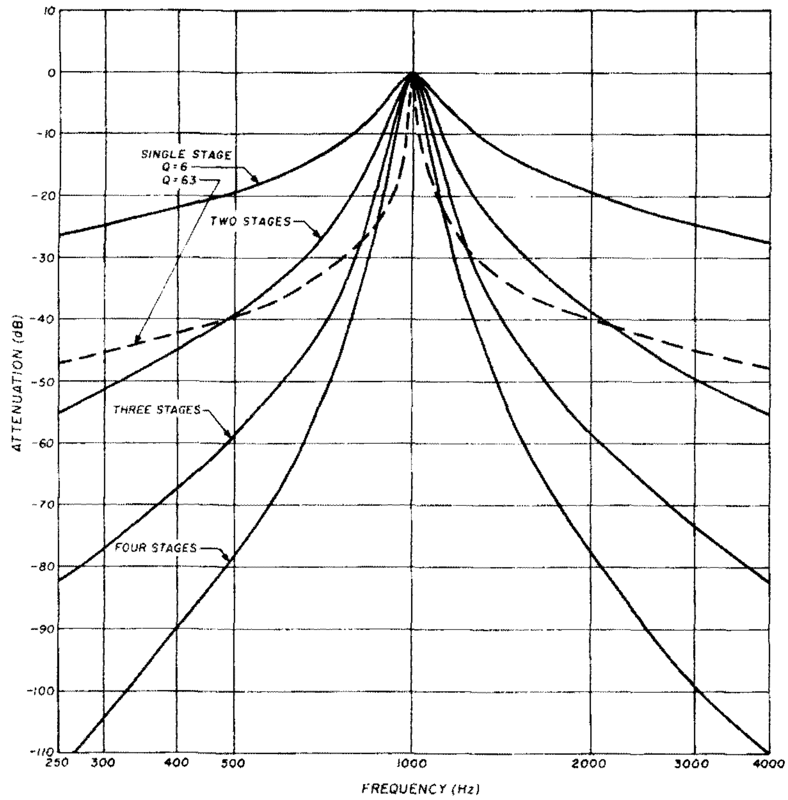


fig. 2. Response characteristics of single and cascaded filter sections. Cascading offers a more practical approach for R1 and R2 values, which are used to vary filter gain and Q (see fig. 1).

Still another circuit that is probably a little more complex is shown in fig. 1. The response and design equations for this filter are:

$$\frac{e_o}{e_i}(\omega) = -\frac{j\omega[1/R1C1]}{\left[\frac{1}{R3R6C1C2}\left(\frac{R5}{R4}\right) - \omega^2\right] + j\omega[1/R2C1]}$$

$$f_o = \frac{1}{2\pi}\sqrt{\left(\frac{1}{R3R6C1C2}\right)\cdot\left(\frac{R5}{R4}\right)}$$

$$\frac{e_o}{e_i}(\omega_o) = -\frac{R2}{R1} \qquad (9)$$

Choose $C1 = C2$
$R4 = R5$
$R3 = R6$

$$Q = 2\pi f_o C1 R2$$

Typical component values chosen for this circuit were:

R1 98,000 ohms	R4 10,000 ohms	C1 0.01 μF
R2 98,000 ohms	R5 10,000 ohms	C2 0.01 μF
R3 15,915 ohms	R6 15,915 ohms	f_o 01000 Hz

3-dB BW 71 Hz (four stages); shape factor 13
(See fig. 1 for final values).

The values for R1, R2, R3 were made variable for the first stage, while a potentiometer was used in the following stages for R3. (See fig. 1.) R1 varies the overall filter gain, R2 the Q, and R3 tunes each stage to the same desired center frequency. A value of 50k was more than adequate to cover the tolerance variations and range of interest. The variable pots, R1 and R2, were 500k maximum for the input stage. The remaining stages included the values given above except for R3, which was chosen to be 50k in series with a 10k resistor.

This approach shows no improvement in skirt selectivity over the others and, in the same manner, requires cascaded stages or building blocks as they are sometimes called.[3] However, this circuit has the advantage of not being so component-tolerance sensitive. In fact, the Q can be varied without affecting the gain or center frequency by varying R2; the gain can be varied independently by varying R1; and the center frequency can be varied independently by varying R6 or R3. The filter can be easily tuned using components with 10-percent tolerance. Also, you can adjust the tone or output level without varying other filter parameters, such as frequency or Q.

This completes the basic description of the three more popular bandpass filter circuits. The following discussion addresses the breadboard testing of each circuit, with a final filter design using the circuit having independent characteristic control.

filter-circuit breadboards

The parallel-T network was evaluated first. The feed-

back network using a parallel-T network was very sensitive to component variation. The measured filter characteristics agreed with theoretical results, using a programmable hand calculator, almost to the point where one would be satisfied without evaluating a breadboard circuit (if 1-percent or better component tolerances were used).

The second circuit evaluated was that used in the popular MWJ-CWF2, which uses matched components for each section — difficult to implement unless you have an impedance bridge. The measured results again agreed with theoretically derived results when 1-percent component tolerances were used.

When reviewing the references on this subject, I found that all attained the same success between theoretical and measured results at audio frequencies when using close tolerance values.[4] As in the parallel-T circuit, the second circuit was sensitive to component-value variation and only slight improvement was obtained regarding characteristic control.

The third circuit was by far the most superior in terms of varying filter characteristics independently of each other. Again, theoretical results agreed with measured results after tuning each stage to the desired center frequency. In all cases, the measured results departed from those of the theoretical case only when the operational amplifier characteristics no longer were allowed to assume a very high input impedance, a near-zero output impedance (power limited of course), and a very high open-loop gain (with respect to the Q and skirt selectivity plus stage-to-stage isolation).

construction and cost

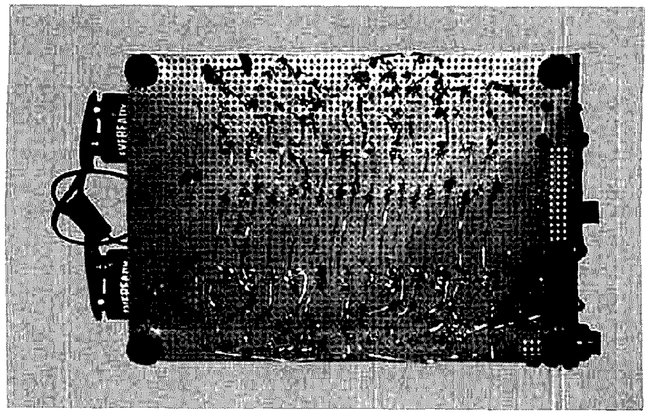
Cost was kept to a minimum by using available barrel-kit components. In this case, 14 good μ741 ICs were found among fifty purchased, for a little less than two dollars. The quad operational amplifier, LM324, although a little more expensive, provided a much neater layout and also showed a slightly higher open-loop gain at 1000 Hz, where the μ741 drops of quite a bit.

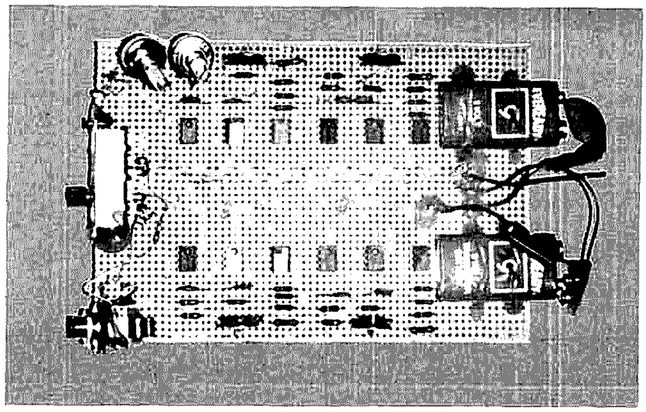
Construction and layout were arranged so that either the quad operational amplifier chips or the single μ741 could be evaluated. The photos show top and bottom component layout. Printed circuit boards could have been used, but construction time and cost would have increased. Batteries were used; however, a separate power supply might be more desirable. The filter draws only about 17 mA.

concluding remarks

Fig. 2 illustrates the response characteristics as the stages are cascaded. At first I thought that one stage with a higher Q would suffice (the dashed curve shows the response characteristic of a single stage with a Q of 63). This arrangement didn't provide the desired skirt attenuation for rejecting strong signals close to center frequency, so additional stages were cascaded to improve the shape factor (60-dB bandwidth divided by the 3-dB bandwidth).

Cascading stages offered more practical and less component-sensitive values for R1 and R2. An important observation noted from the response curves is that only one stage needs a variable pot for varying Q, at least within the ranges normally desired. If a greater Q range is desired, pots can be used in all four sections.



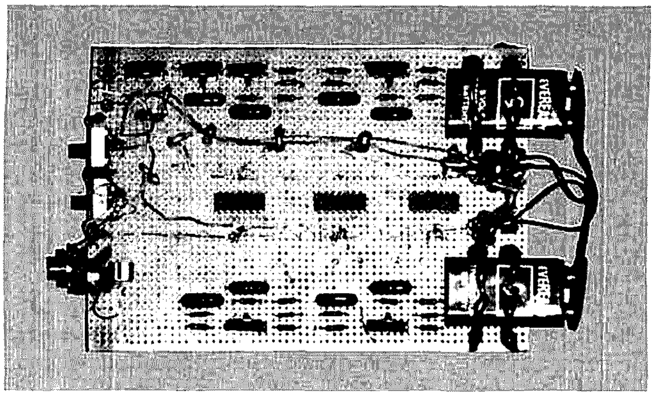


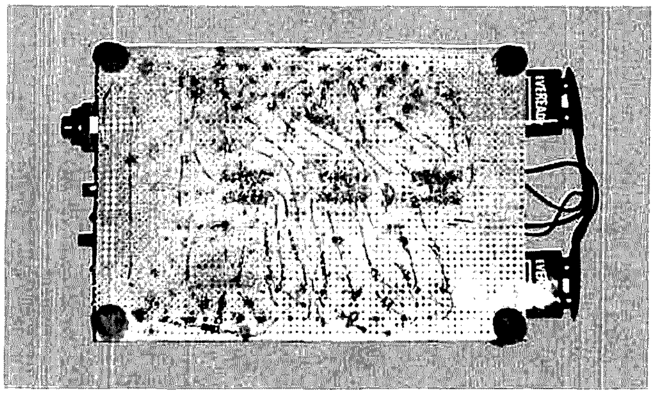
Top and bottom (above and below) of the Vector board construction used for the experimental active bandpass filter using a single μ741 op amp.

The more practical approach for wide Q variation is to use a four-stage pot with a common shaft coupling. Similarly, the frequency adjustment requires only one pot per stage unless a wide frequency range is desired. Normally R5 is chosen to equal R4; C1 to equal C2; and R3 to equal R6 for maximum dynamic range. However, varying only one of the resistors, R3 or R6, seems to allow ample dynamic range, especially if the filter is installed immediately after the detector in a receiver. A center frequency between 700 and 1500 Hz is usually chosen.

It will also be noted from **fig. 2** that the response is not affected nearly as much near the center frequency when cascading stages as it is far removed from the center frequency. This further illustrates the need for cascading stages to obtain the desired skirt selectivity.

The two circuits shown in **fig. 1** illustrate the input stage, A, which includes the pots for Q and gain adjustment and the following additional cascaded stages, B, each of which uses a pot for tuning to center frequency. The op amps, with their near-infinite input impedance





The experimental quad op amp active bandpass filter showing component layout, above, and wiring, below, using Vector board construction for flexibility and low cost.

and near-zero output impedance, provided excellent isolation between filter stages in the feedback loop. In this case, the noninverting inputs were connected to ground along with the center common point of the two bias sources of plus and minus 9 volts.

Very weak signals could be pulled out of the noise by increasing the first-stage filter gain and Q. The filter exhibits an increase in weak-signal level when the gain and Q are adjusted carefully by presenting energy near the center frequency of the filter, which causes it to ring slightly. Filter ringing increases the desired signal output level and provides an apparent signal-processing gain, which strengthens the lower signal voltage levels. This does not mean that energy is created, but that energy is added to the amount detected.

It was not the intent of this article to describe each filter in great detail but rather to present an introduction and the supporting design equations. Many other configurations are possible, but I feel that those discussed here are of primary interest. At any rate, you can use the design equations for your own audio bandpass filter and tailor it to your requirements.

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2. J.G. Graeme, G.E. Tobey, and L.P. Huelsman, *Operational Amplifier Design and Applications*, 1973 Burr-Brown Electronic Series, McGraw-Hill, New York.
3. P.E. Fleischer and J. Tow, "Design Formulas for Biquad Active Filters Using Three Operational Amplifiers," *Proceedings of the IEEE* (letter), May, 1973, page 662.
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ham radio

ten commandments for technicians

I Beware the lightning that lurketh in the undischarged capacitor, lest it cause thee to bounce upon thy buttocks in a most untechnician-like manner.

II Cause thou the switch that supplieth large quantities of juice to be opened and thusly tagged, that thy days in this earthly veil of tears may be long.

III Prove to thyself that all circuits that radiateth and upon which thou worketh are grounded and thusly tagged lest they lift thee to radio frequency potential and causeth thee also to make like a radiator.

IV Tarry not amongst those fools who engageth in intentional shocks for they are surely nonbelievers and are not long for this world.

V Take care that thou useth the proper method when thou takest the measure of a high-voltage circuit lest thou incinerate both thyself and thy meter, for verily, though thou hast no account number and can easily be surveyed, the test meter doth have one and, as a consequence, bringeth much woe unto the supply department.

VI Take care that thou tampereth not with safety devices and interlocks, for this incurreth the wrath of thy supervisor and bringeth the fury of thy safety inspector down upon thy head.

VII Work thou not on energized equipment, for if thou dost, thy fellow workers will surely buy beers for thy widow and console her in other ways.

VIII Service thou not equipment for electrical cooking. It is a slothful process and thou might sizzle in thine own fat for hours upon a hot circuit before thy Maker sees fit to end thy misery.

IX Trifle thou not with radioactive tubes and substances lest thou commence to glow in the dark like a lightning bug and thy wife have no further use for thee except thy wages.

X Thou shalt not make unauthorized modifications to equipment, but causeth thou to record all field changes and authorized modifications made by thee, lest thy successor tear his hair out and go slowly mad in his attempt to decide what manner of creature hath made a nest in the wiring of such equipment.

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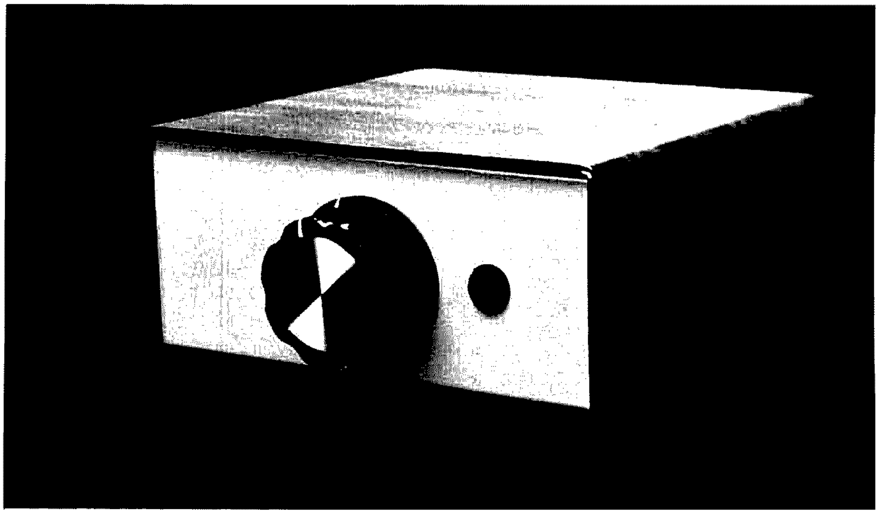


wideband preamp

Construction details for a low noise, wideband receiver preamp that covers the frequency range from 1000 kHz to 50 MHz

Here's an effective yet inexpensive wideband preamp that covers the amateur high-frequency bands. Use it to soup up old receivers of the pre-1970 era or improve performance of more modern equipment. The circuit is easy to build, requires no exotic parts, and should go together in one or two evenings. Total parts cost is less than ten dollars.

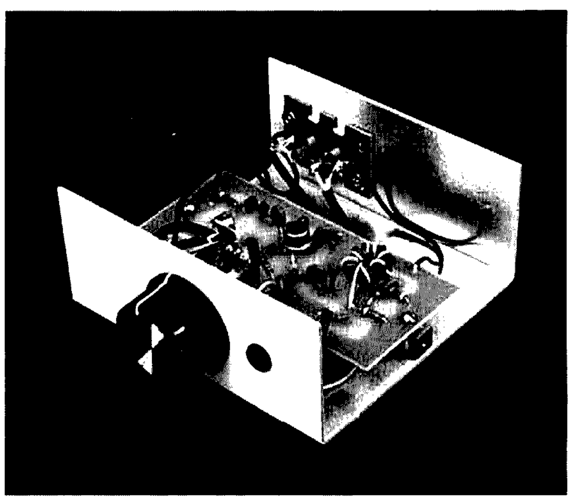
The circuit (fig. 1) features a 2N5109 npn silicon high-frequency transistor, which is used in CATV amplifiers requiring low cross-modulation distortion and low-noise input. The 2N5109 has a 3-dB noise figure at 200 MHz, current gain-bandwidth product of 1200 MHz, cross-modulation distortion of −70dB, and collector-emitter amplifier voltage gain of 11 dB (50-216 MHz). It sells for about $3.00 in small quantities. A PC-board layout is shown in fig. 2.

While the circuit was intended to be used in the frequency range from 1 to 50 MHz, the 2N5109 is a hot performer well into the vhf range. However, the coupling and bypass capacitors, balun transformer, and circuit layout may need to be changed to reach the higher frequencies. In certain cases, as when using a wideband antenna or equipment with poor intermodulation or spurious-response characteristics, it may be necessary to use bandpass filters[1] on the amplifier input or output.

By Ed Pacyna, W1AAZ, Danbury Circle, Amherst, New Hampshire 03031

acknowledgement

I would like to acknowledge the contributions of Jim Jenkins, K1JXK, who's early ideas led to my writing this article.



The printed-circuit board for the wideband rf preamplifier is built into a 4x4x1½ inch (10x10x3.8cm) aluminum enclosure. LED is mounted inside rubber grommet to the right of the in/out switch knob.

reference

1. Wes Hayward, W7ZOI, "Bandpass Filters for Receiver Preselectors," *ham radio*, February, 1975, page 18.

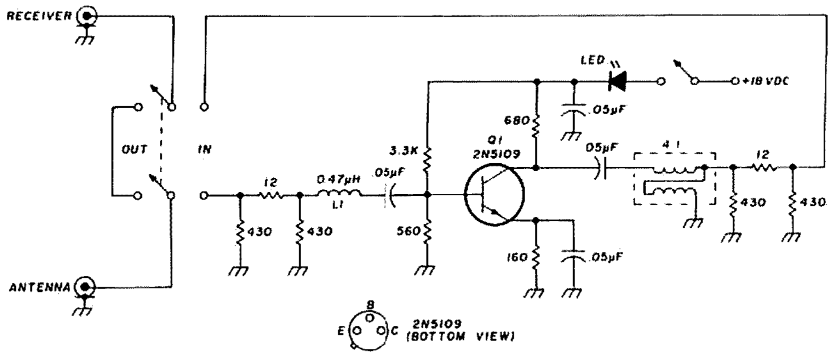


fig. 1. Wideband preamp schematic. The .47 µH inductor can be a couple of ferrite beads placed over a short loop of no. 20 to no. 30 AWG (0.8-0.25mm) insulated wire. The 4:1 wideband transformer in the transistor collector circuit is made by twisting together two pieces of insulated 20 to 30 AWG (0.8-0.25mm) wire, then winding 8 to 10 bifilar turns on a ¼ to ½ inch (6-13mm) toroid core.

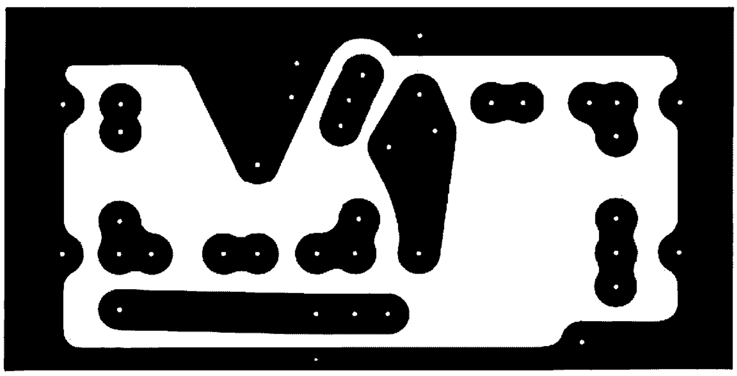


fig. 2. Foil side of the printed-circuit layout for the wideband amplifier.

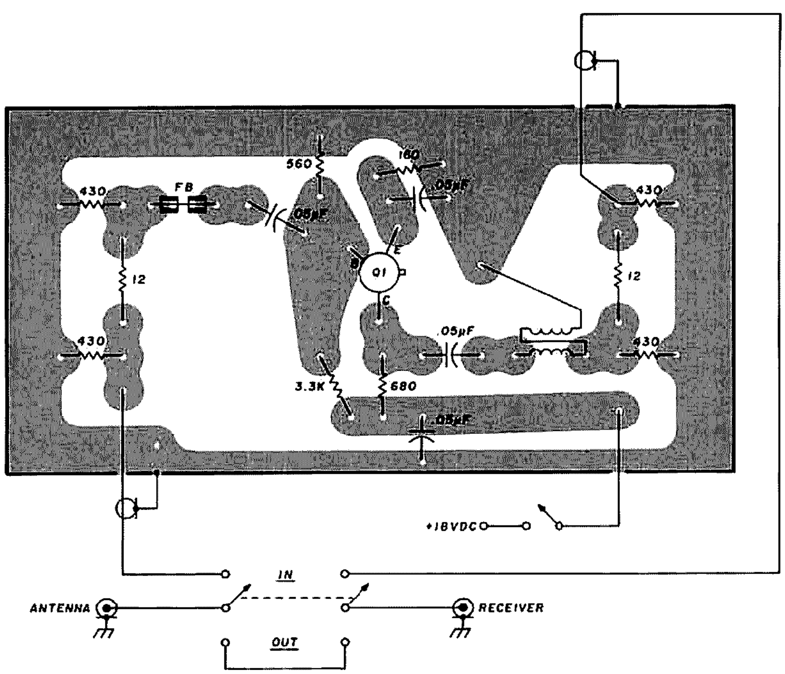


fig. 3. Component layout for the wideband high-frequency preamplifier.

four-band
vhf receiving converter

One 46-MHz crystal and an ingenious mixing scheme provide coverage between six meters and 420 MHz

How would you like an inexpensive way to monitor your local two-meter repeater? Or how about a 220-MHz receiver? In fact, how about a receiver covering 6 meters through 420 MHz? These are a few of the questions that were buzzing around in my head, so I decided to build a general-purpose receiver capable of covering all vhf bands. The result is the front end of an all-band vhf converter.

design

At a recent test-equipment seminar, a nifty but little-used fact concerning mixers surfaced once again: Not only is the local oscillator's fundamental frequency available; but the mixer itself, being a nonlinear device, produces harmonics of the oscillator frequency that can be used in the mixing action.

Example: Use a 46-MHz local oscillator and an input of 51 MHz to generate an i-f at 5 MHz. The local oscillator's 138-MHz third harmonic can produce the same i-f from an input at 143 MHz. In fact, this i-f will be produced from any of the following input frequencies: 87 MHz, 97 MHz (2 x 46 = 92); 133 MHz, 143 MHz (3 x 46 = 138); 179 MHz (4 x 46 = 184); 179 MHz, 189 MHz (4 x 46 = 184); and so on.

Now, 46 MHz is not just a number plucked out of thin air for this example. It was developed after weeks of doodling, and is a key ingredient in the design. To understand why, consider this:

1 x 46 = 46	(used on 6 meters)
3 x 46 = 138	(used on 2 meters)
5 x 46 = 230	(used for 220 MHz)
7 x 46 = 322	(not used)
9 x 46 = 414	(used for 420 MHz)

Therefore, it should be possible to use only one crystal and some sort of odd-harmonic generator to meet the design objective of all-vhf-band coverage. 420 MHz is picked up as a bonus (vhf ends at 300 MHz).

circuit description

Fig. 1 is the complete converter schematic with parts list. Ignoring the broadband rf preamplifier (Q1 through Q3) for the moment, let us continue our study of the harmonic mixer. Crystal Y1, a plated, overtone unit, oscillates at 46 MHz in a series-resonant mode. Transistors Q5 and Q6 form the differential amplifier oscillator: rf voltage at Q5's collector is fed back to Q6's base by capacitor C5 and the crystal.

Harmonic emphasis is supplied by tuned line L1 connected from Q6's base to rf ground. L1 is a shorted, quarter-wave transmission line cut to resonance at the crystal frequency. It is 3/4 wavelength long at the third harmonic, 5/4 wavelength long at the fifth harmonic, etc., and acts as a parallel-tuned circuit at all odd harmonics. Therefore, the fundamental and odd harmonics of the crystal are present at the base of mixer transistor Q6, as desired. Construction details for a miniaturized quarter-wave line suitable for PC-board mounting are given in fig. 2, which is self-explanatory.

The remainder of the converter consists of a broad-

By S. Smith, W3TQM, Laurel, Maryland 20810

band, cascode preamplifier (transistors Q1, Q2, and Q3) that is R-C coupled to mixer driver Q4. No tuning is incorporated purposely so that the converter may be conveniently used on any of the four bands. Of course, external tuning is required, or the device will work on all

to figs. 2 and 3, you should have no difficulty building one of these converters.

The six transistors, in two differential amplifier configurations, are packaged as RCA's CA3049T integrated circuit. This IC was reportedly designed for low-noise

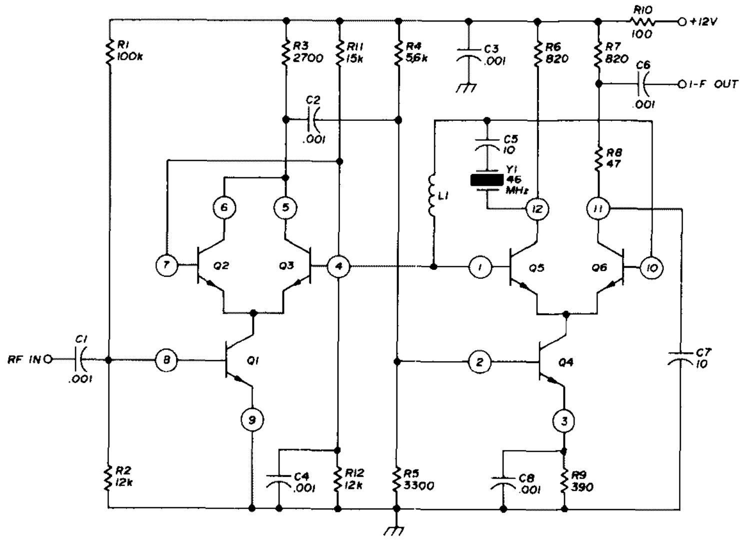


fig. 1. Schematic of the four-band vhf receiving converter. Resistors are ½ watt, 10% composition. All transitors are part of RCA CA3049T integrated circuit. Circled numbers are IC pin numbers. Capacitor voltage rating is 25 volts working or greater. Construction of transmission line resonator, L1, is shown in fig. 2.

bands at once! It also converts local TV and fm frequencies, because even harmonics of the crystal are not eliminated in the design.

construction

Fig. 3A is the foil pattern I used. A glass/epoxy board, 2 by 2 inches (51x51mm) is used. Fig. 3B is a parts layout diagram. Note that all components mount on top of the board; the foil is the bottom. By reference

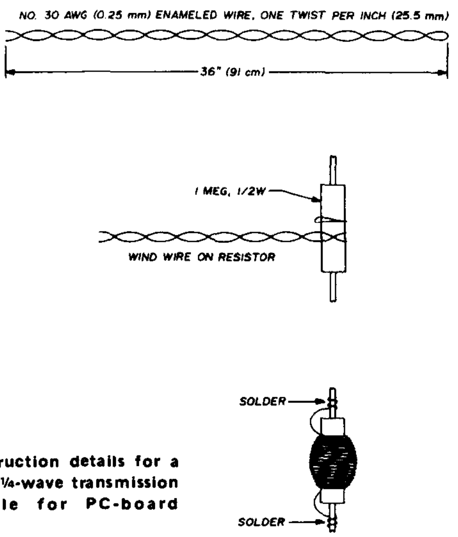


fig. 2. Construction details for a miniaturized ¼-wave transmission line suitable for PC-board mounting.

vhf amplifier service and was chosen for its small size and cost.

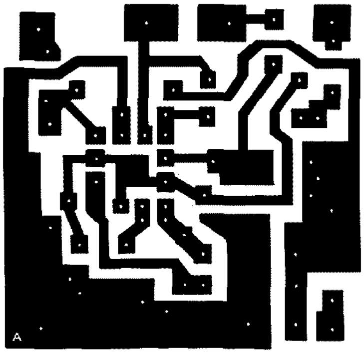
performance

Several converters I built had sensitivities below 1 μV on 50, 144, and 220 MHz, and sensitivities of about 2 μV on 420 MHz. Image rejection and i-f feedthrough, being functions of the input tuning, are not meaningfully measured on such a broadband device. These parameters vary according to the tuning scheme used. Fig. 4 is a simple tuner that serves the purpose and matches the converter input to a 50-ohm antenna. Note that, with a bit of care, this input tuner could be bandswitched.

Conversion gain is a function of the band in use. The output level obtained during sensitivity checks, at 10 dB s/n ratio, is 10 μV. Power required is 12 Vdc at approximately 15 mA; the circuit works well on a 9-volt supply with slightly reduced sensitivity.

applications

I use one converter to monitor 146.76 and 146.94 MHz on a broadcast-band transistor radio. To do this, the crystal was changed to 48.5 MHz, achieving an injection frequency of 3 x 48.5 = 145.5 MHz. The resulting i-f is in the broadcast band, and the segment of two meters between 146 and 147 MHz can be tuned by tuning the broadcast receiver through the i-f range of 0.5 to 1.5 MHz. For 146.76 MHz, the i-f converts to 1260 kHz, while 146.94 MHz converts to 1440 kHz. Another common vhf fm frequency, 146.52, converts to 1020 kHz.



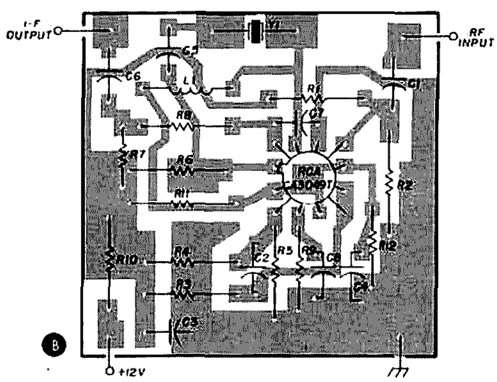


fig. 3. Foil side of the PC board, A, and parts layout, B. Note that all components are mounted on top of the board; foil is the bottom.

The i-f output can be inductively coupled to a pocket transistor radio or directly coupled to an auto broadcast receiver. See fig. 5 for suggested coupling methods. In either case, fm can be recovered quite successfully by slope detection. When using a pocket transistor set, interference from broadcast stations can be removed by wrapping the radio in aluminum foil. Be sure to put the rf choke of fig. 5 *inside* the foil and as close as possible to the radio's Loop-Stick antenna.

Another of the converters is used to monitor the local repeater output at 146.76 MHz on my Drake R-4A receiver. Again, a crystal other than 46 MHz is used, so that the i-f can be tuned with the receiver. In this case, a 46.5-MHz crystal yields an i-f of 7.460 MHz, which is in range of the Drake's 40 meter band. Once again, slope detection is used to recover fm; some distortion is experienced because of the receiver's selectivity, but no trouble is experienced copying traffic.

Yet another unit is in use converting both 15- and 10-meter signals to an i-f in the range of a surplus BC-348 receiver. A crystal frequency of 25 MHz was chosen. It was necessary to replace the tuned line, L1, with a parallel-resonant circuit at the crystal frequency.

The converter may also be used as a general-coverage vhf/uhf receiver. Any i-f up to 35 MHz may be used without modification of the original circuit, and other intermediate frequencies could be adapted by modification of the output stage.

As all crystal harmonics are present in the mixer, a frequency in the 25 to 1000 MHz range can be converted to an i-f of less than 35 MHz. As an example, using the second harmonic of 46 MHz (92 MHz), 121.5 MHz would be converted to 29.5 MHz. Using the third harmonic (138 MHz), this same signal would be found at 16.5 MHz. Thus, the converter can be used to monitor aviation, police, business, and other frequencies. Again, the tuned input circuit should be resonated to the desired frequency. Problems experienced when the signal frequency falls close to or on a crystal harmonic can be circumvented by using a second crystal differing in frequency from the 46-MHz standard, so that low i-fs are avoided.

Now there's little excuse for not monitoring the higher-frequency amateur bands. This project can be built in a few evenings, and for less than $20.00. It's inexpensive, simple, and it works. See you on vhf.

ham radio

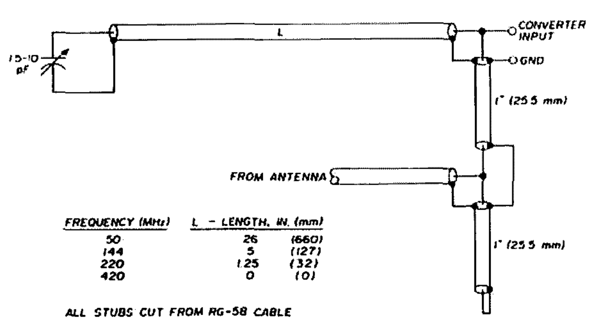


fig. 4. Simple tuner for reducing feedthrough and images.

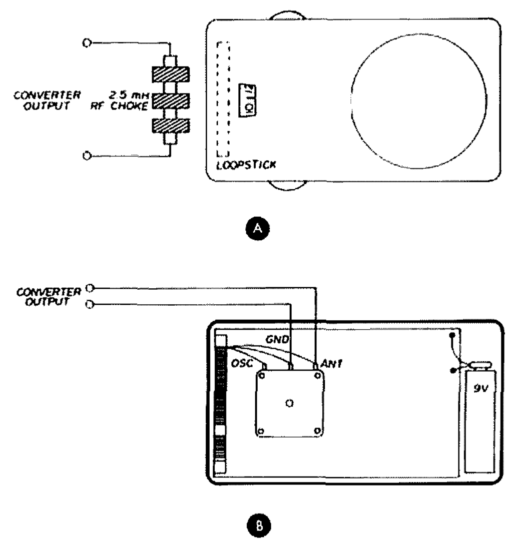


fig. 5. Methods for coupling the vhf converter to a pocket transistor broadcast-band radio, A, or an automobile broadcast-band set, B. Transistor radio and rf choke should be wrapped in aluminum foil to eliminate BC-band interference.

phasing-type single-signal detector

Phase-shift network design and construction data for improving receivers with marginal selectivity

Phasing-type detectors, despite their superficial complexity, are the only means of avoiding audio images in a direct-conversion receiver. Likewise if a superheterodyne receiver has inadequate selectivity or is plagued by broad-band i-f noise,[1,2] the cure lies in either additional i-f filters or in a phasing detector.

Although the circuits described here would make a good direct-conversion receiver, they were designed for use in a superhet with lots of i-f noise and inadequate stop-band rejection. While this system can't offer all the advantages of a second crystal filter, it does give substantial improvement at moderate cost. (Phase-shift networks are described in the appendix for unwanted sideband suppressions of at least 30, 37, and 60 dB, depending on the audio bandwidth and circuit complexity. A simple network for CW reception offering 50 dB rejection is also described).

The circuits were chosen for their ease of construction and adjustment. No special LCR precision bridges are needed to set the audio phase network capacitors. If

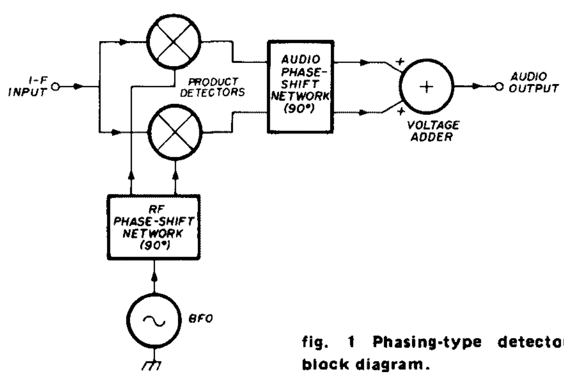


fig. 1 Phasing-type detector block diagram.

you can tune a musical instrument the only test equipment needed is a grid-dip oscillator and a vtvm. A frequency counter and oscilloscope are useful but not absolutely required. The system costs between $5 and $25 more than a simple bfo-product detector combination depending on how good a scrounger you are. (Compare this price with the cost of crystal filters!) The phasing detector block diagram is shown in **fig. 1**.

By Don Lawson, WB9CYY, 4601 Jay Drive, Madison, Wisconsin 53704

The audio phase-shift networks described here are also suitable for use in ssb phasing-type transmitters. For such applications wire the two inputs (C and D of **fig. 3**) together and run the outputs (TP1 and TP2) to the balanced modulators.

bfo and product detector

The bfo and product detector schematic is given in **fig. 2**. L2 and C2 resonate at the bfo frequency; R1 is should be chosen so that the combined drain current of both mosfets is equal to the average of their individual values of I_{Dss} so that about 9 volts are between the source and drain of each transistor (whichever condition gives the larger value of R4).

R2 and R3 can have values anywhere between 470 ohms and 1k and should be closely matched. C4 and C5 should also be matched to 5% or so (using the method described later). L3, L4, C6, and C7 form a lowpass

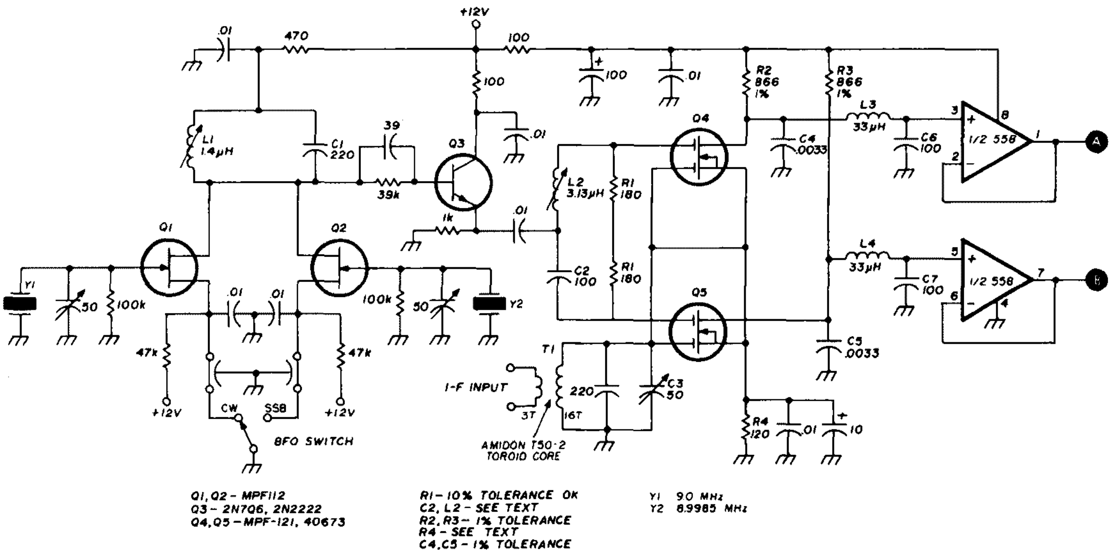


fig. 2. Bfo-product detector schematic. Components that must be closely matched are discussed in the text.

equal to the reactance of the capacitor or inductor at resonance. For an i-f of 9 MHz, representative values are R1 = 180, C2 = 100 pF, and L2 = 3.13 μH.

The bfo uses about the simplest method I know for switching parallel-resonant crystals, and with MPF-112s at 35 cents apiece, it's fairly inexpensive. The product detector's dual gate mosfets, Q4 and Q5, can be MPF-121s or 40673s. If you buy three of four mosfets (instead of just two), you can be fairly sure of finding two that have similar values of V_{gsoff} and I_{Dss}. R4

filter to prevent the bfo signal from getting into the op-amps. The bfo-product detector assembly should be built into a shielded enclosure to keep the bfo signal out of the i-f strip.

audio circuits

The audio phase-shift network is shown in **fig. 3**. There are simpler circuits in the literature but this one is by far the easiest to build. C1, C2, C3, and C4 should be set to equal values by padding with small capacitors in

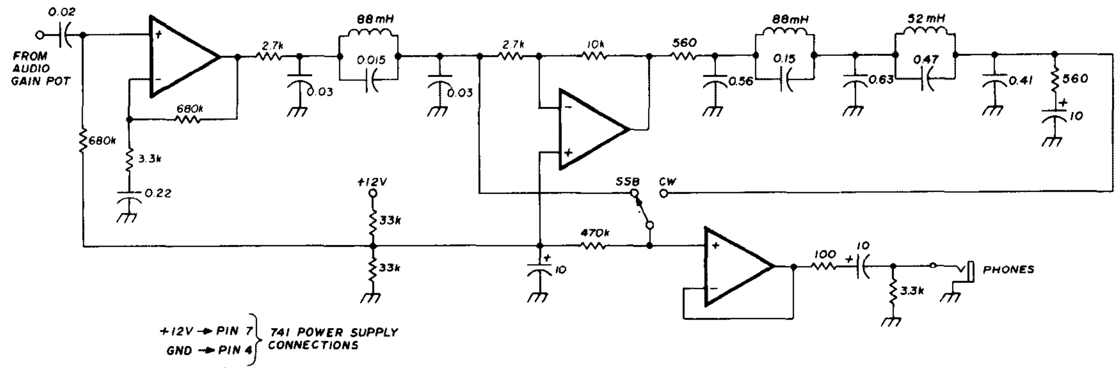


fig. 4. Audio amplifier and filter schematic. Cutoff frequencies are respectively 3 kHz and 800 Hz for the single- and double-section filters.

parallel. Several 50, 100, and 200 pF capacitors are handy for this. If you want to use some other value for the capacitors, multiply R1 through R4 by your capacitor value divided by 0.022; i.e., for 0.1 μF, R1 would be 23.03k x 0.1/0.022 or 104.7k. The phase-shift network is designed to work from 300 to 4000 Hz and theoretically provides at least 31 dB suppression. For this circuit, R5 through R8 form voltage dividers. In the circuit of fig. 3 they have a ratio of resistances of 14.94 to 1.

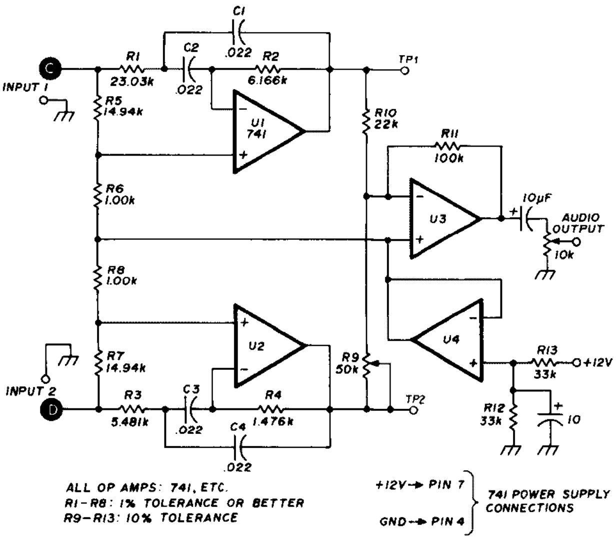


fig. 3. Audio phase-shift network schematic. C1-C4 should be matched by adding small capacitors in parallel.

R9, R10, R11, and U3 form the voltage adder of fig. 1. U4 gives a good ac ground for the networks. Trying to substitute a large bypass capacitor for U4 isn't good enough. To eliminate U4 you'd need dual power supplies for U1 through U3, so that dc ground is also a suitable ac ground. The audio output can be applied to an amplifier with sharp filters; a typical circuit is shown in fig. 4. The single-section filter is a 3-kHz elliptical lowpass; the two-section filter has a cutoff frequency of 800 Hz for CW reception.

construction and alignment

A grid-dip oscillator is all that's necessary for building and checking the bfo and product detector. If an oscilloscope is available, it might be a good idea at this point to set L2 initially for the point where the signals at A and B (fig. 2) are 90 degrees apart in phase. (Use the circuit of fig. 5A if a scope with a triggered sweep is available; use fig. 5B if one is not).

All capacitors in the audio phase-shift network should be matched as closely as possible and R1-R4 should be proportional to their calculated values. The ratios of R5/R6; R7/R8 should also be as close as possible to their computed ratios. For example, if all your capacitors were equal to 0.019 μF and R1-R4 were 0.9 times their computed values, the only bad effect would be that

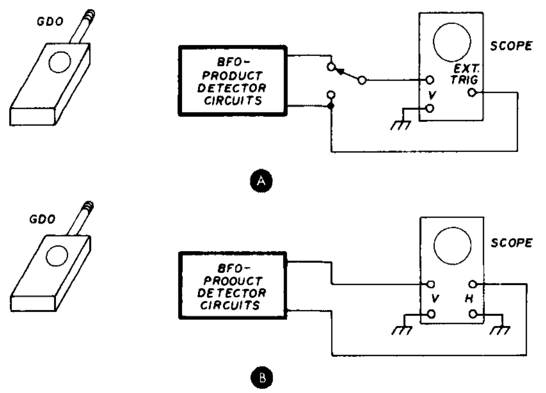


fig. 5. Setups for checking and presetting the rf phase-shift network. Methods (A) and (B) are for oscilloscopes with and without triggered sweep.

the network frequency range would be shifted slightly. The resistors can be made up from parts on hand, using an accurate volt-ohmmeter, or 1% wirewound resistors may be used.

The capacitors can be matched quite well by using the oscillator circuit of fig. 6. Be sure to use an uncom-

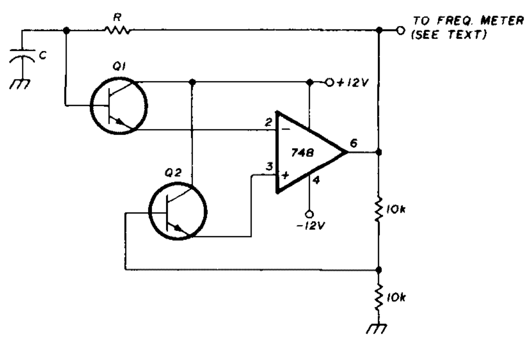


fig. 6. Oscillator circuit for measuring capacitance. An uncompensated op-amp should be used. Q1, Q2 are general-purpose audio transistors.

pensated op-amp, such as a 748 or a 709. The internal compensation in a 741 will slow down the oscillator and measurements will be unreliable. Try to keep the frequency between 1 and 10 kHz for best accuracy. If you're going for readings of four significant digits, power the oscillator from either batteries or a regulated supply to eliminate effects of line-voltage variations.

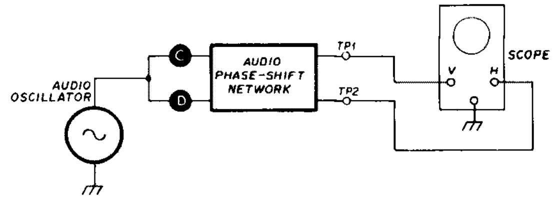


fig. 7. Test circuit for checking the audio frequency phase-shift network.

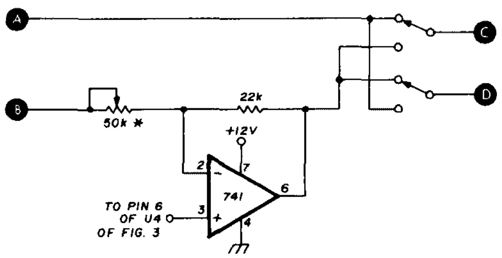


fig. 8. Method for sideband selection through a panel switch. Adjust the 50k pot so that the signal amplitude at pin 6 of the 741 is the same as at point A.

The "frequency meter" should be either a digital frequency counter or a tape recorder. A meter-movement type instrument won't allow sufficient precision. If you use a tape recorder, find which capacitor gives the lowest-pitch tone then pad the other capacitors to give this same frequency. By taping the lowest pitch, you'll save having to switch constantly between capacitors for comparison. If you use a counter, choose one capacitor as a reference and note its frequency. Assuming it equals its marked value exactly, the relative capacitance of the other capacitor is:

$$C_x = \frac{freq}{f_x} \ x \ C_{ref}$$

Unless you're using a very good ohmmeter to measure the resistors, there's little point in matching the capacitors to better than 0.25%. If possible, use high-stability capacitors (polystyrene is best).* If you're using carbon-composition resistors, coat them with epoxy to reduce

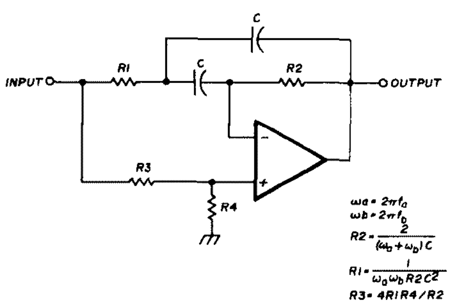


fig. 9. Portion of a phase-shift network with singular frequencies, f_a and f_b. Formulas are used with table 1 to compute component values.

variations of their values with humidity. Use fig. 7 to check the operation of the phase-shift network. With the scope setup shown you should be able to see a circle on the scope screen.

Referring to figs. 2 and 3, in final adjustment first wire A to C and B to D. Sweep the grid-dip oscillator or signal generator past the bfo frequency and decide which sideband (upper or lower) is being rejected. If it's not the sideband you want, try wiring A to D and B to C. (If you want to select sidebands by a panel switch, try using the circuit of fig. 8). Now, alternatively adjust L2 of the bfo and R9 of the audio phase-shift network for the best unwanted sideband rejection.

results

If you use the circuit in fig. 3, you should be able to easily get about 30 dB of unwanted sideband rejection between 300-4000 Hz. For better rejection, use one of the other circuits in the appendix.

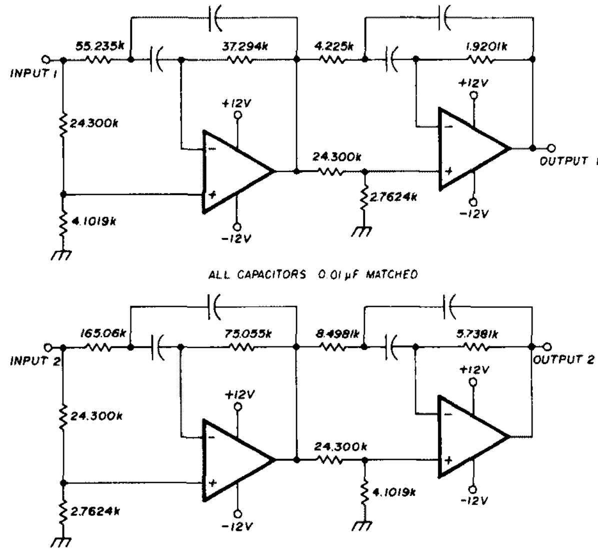


fig. 10. Cascaded filters to obtain an 8-pole network. Circuit provides at least 60 dB unwanted sideband rejection between 200-4000 Hz using 741 op-amps.

I hope this article provides some ideas for those who build or modify receivers. When it comes to improving the selectivity of a marginal receiver, this system is the best for the money.

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*Polystyrene capacitors can be obtained from Weinschenker, Box 353, Irwin, Pennsylvania 15642.

appendix

The design of any phase-shift network is a trade between complexity, bandwidth, and phase accuracy. Phase-shift networks are a special category of "all-pass filters," meaning that the network gain is constant regardless of frequency. Phase-shift networks are characterized by a set of "singular frequencies." The calculation of these frequencies is too involved to be covered here, but an algorithm is in the literature that requires nothing more than a calculator with a square-root key.[3,4] For those who don't like lots of arithmetic, I've compiled the results of several design computations in table 1. Network 1 is the general-purpose network shown in fig. 3, while network 2 is for more narrow-band ssb use. Network 3 is for use in direct-conversion CW receivers. These networks are all of the four-pole variety since each has four singular frequencies.

table 1. Phase-shift network design data.

network	freq range (Hz)	minimum unwanted sideband rejection (dB)	singular frequencies f_1	f_2	f_3	f_4
1	300-4000	31	148.9	625.8	1916.	8049.
2	300-2500	37	136.2	525.2	1428.	5507.
3	225-950	49	88.7	298.6	672.0	2265

As noted before, the circuit used here was chosen for its simplicity of construction as compared to other circuits.[5] (Unless you have a frequency counter or precision LCR bridge, it's much easier to set capacitors equal than it is to set them to arbitrary ratios). To compute component values from table 1, we take either odd- or even-numbered singular frequencies and plug them into the formulas of fig. 9.

As an example, let's try designing the network shown in fig. 3. As a start, we take odd frequencies of 148.9 and 1916 Hz. Converting to radians/sec, we get $\omega_1 = 935.6$ and $\omega_3 = 12{,}038$. Assuming $C = 0.025\ \mu F\ (2.5 \times 10^{-8}\ F)$, we compute

$$R2 = \frac{2}{(935.6 + 12{,}038) \times 2.5 \times 10^{-8}} = 6166\ ohms$$

Likewise,

$$R1 = \frac{1}{935.6 \times 12{,}038 \times (2.5 \times 10^{-8})^2 \times 6166}$$
$$= 23.040\ ohms = 23.04k$$

Finally, since we have a couple of 1k 1% resistors on hand, we choose R4 = 1k and

$$R3 = \frac{4 \times R1 \times R4}{R2} = \frac{4 \times 23{,}040 \times 1000}{6166}$$
$$= 14.946\ ohms = 14.95k$$

We now have half of our network designed. For the other half we follow the same procedure except that we work with f_2 and f_4.

Note that at dc the circuit gain is $R4/(R3 + R4)$. Since this is an all-pass network, all audio frequencies are attenuated by this same amount. Hence, this network attenuates the signal by 24 dB. Because the 741 op amps have internal noise it may be desirable, especially in direct-conversion receivers, to put low-noise amplifiers (such as an LM381) between the product detector and the phase-shift network.

A final note: It's possible to cascade phase-shift network sections to get an 8-pole network. A practical 8-pole network is shown in fig. 10. According to a computer simulation, this circuit provides at least 60 dB of unwanted sideband rejection between 200-4000 Hz. This simulation was with 741s. It may be possible to get an additional 5 dB suppression by using 556 op-amps since their input impedance is much higher than that of a 741. Of course, the components would have to be set very accurately to realize this theoretical figure and a preamplifier would be necessary to overcome the network signal attenuation. This circuit would be ideal for a high-quality, phasing-type ssb direct-conversion receiver.

ham radio

the ham notebook

another squelch circuit

Interest has developed in the squelch circuit appearing in the September, 1974, issue of *ham radio*, page 68, which was published before the inexpensive and popular LM380 audio amplifier IC became readily available. Fig. 1 shows a squelch circuit with excellent performance and greater simplicity.*

Transistor Q1 amplifies random noise which is greater in frequency than the normal spectrum occupied by voice communications during no-signal conditions. Diodes CR1 and CR2 rectify the noise, which is filtered by R6, C5. Q2 conducts and clamps U1 off during no-signal conditions. A carrier activates the LM380. R6 determines the length of the squelch tail, and R5 controls the promptness to which the circuit responds to carrier signals. Increasing the value of C1 increases the noise-amplifier gain, but the smaller the value of C1, the less susceptible the circuit will be in responding to heavy voice peaks, thus the more positive its operation. Cost of the parts at the time of this writing: Q1 \$0.45, Q2 \$0.77, U1 \$1.65. Not bad!

Robert Harris, WB4WSU

*A complete kit, including circuit board, of the audio amplifier section only (less volume control and speaker) is available for \$4.25 from Tekpro Design Systems, Box 6324, Virginia Beach, Virginia 23456.

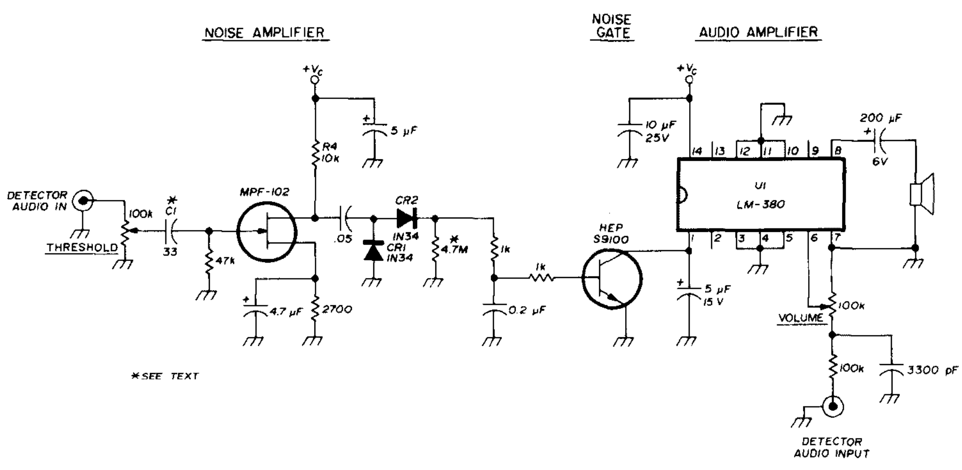


fig. 1. Squelch circuit using the M-380 audio amplifier IC.

transfer letters

In the past I have had great difficulty with dry transfer letters failing to adhere to various surfaces, especially unfinished aluminum. There are two techniques which I have found helpful in overcoming this problem. The first is to abrade the surface to be lettered gently with very fine (0000 or finer) sandpaper and water, and then to dry it thoroughly before lettering. Alternatively, swab the area to be lettered with acetone (most nail polish removers will do). Other solvents probably would work as well, but I have not experimented with them. Either of these methods should always be used before trying to letter a raw aluminum surface that has been handled, or etched in a sodium hydroxide bath (this process leaves a waxy film on the aluminum, and the letters will not adhere to it).

Michael Tortorella, WA2TGL

voltage safety valve

The safety circuit in fig. 2 was developed after several accidents where a power transistor shorted in a low-

voltage power supply causing the transformer output voltage to be supplied to a solid-state transceiver. This voltage could be up to three times the rectified voltage and could cause considerable damage to the transceiver.

A high wattage zener diode could be used across the output of the power supplied, but it takes a very large zener to blow a fuse. Furthermore, the zener could burn open if used to carry currents above its rating.

What was needed was a circuit that would respond when a set limiting voltage was reached and cut off the current flow. The circuit in **fig. 2** is simple and cheap, and uses a relay with contacts large enough to handle the required current (6 or 8 amperes should be ample). The relay coil should be about 200 ohms, the potentiometer 100 ohms or more, the zener diode is a 6 or 8 volt, 10 watt unit.

This circuit can be adjusted to trip at 14 to 24 volts. When the limiting voltage is reached the relay opens up, breaking the circuit to the transceiver. The residual charge in the circuit will hold the relay open until the voltage drops or the power is disconnected.

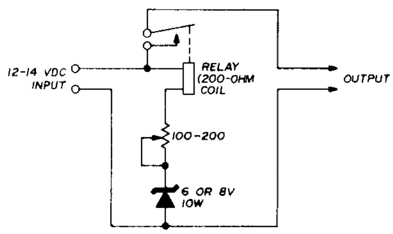


fig. 2. Over-voltage protection circuit. Potentiometer can be set so relay picks up at any preset voltage between 14 and 24 volts.

Harold C. Dressel, W2UVF

Swan 250 carrier suppression

Most Swan 250 transceivers have only about 30 to 35 dB carrier rejection. I have done some work on my unit and now have obtained more than 50 dB carrier rejection. All I did was put 1% resistors in the plate and screen circuits, remove the original 5k carrier balance pot, and install a ten-turn 2000-ohm pot with 1.5k resistors on each side of the pot to give finer control (see circuit in fig. 3).

After doing this I can still insert enough carrier to get full power output from the Swan 250. Before the modification I had about 150 mW of output power with the carrier completely nulled; now, with the carrier completely nulled, output is only 1.5 mW. All measurements were made with a Bird Thruline wattmeter.

Charles A. Beener, WB8LGA

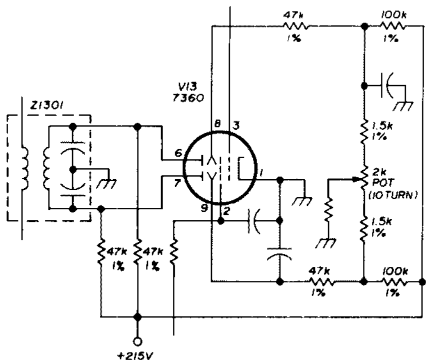


fig. 3. New carrier-balance circuit for the Swan 250 increases carrier suppression to 50 dB or more. The 2000-ohm pot is a miniature, precision 10-turn unit (Bournes 3707).

using your signal generator for absorption measurements

An absorption meter measures the frequency of an rf circuit to which it is coupled. Like a conventional signal generator, it uses an rf tank circuit to tune over the desired band. Therefore, it is logical to combine both functions in a single instrument. I modified my RCA model WR-50A signal generator to also operate as an absorption meter. Performance of the generator is not affected in any way.

The following are added: two RCA-type phono jacks on the front panel, an rf detecting diode, and a pickup (search) coil (see fig. 4). J1 is for the coil, J2 for a 50 μA meter. There is only a slight error in the dial calibration for absorption measurements (due to the pickup coil shunting the cathode portion of the generator tank).

The search coil does not have to be plugged in at the panel where it would be awkard to couple to anything, but may be plugged in at the far end of an extension cable. I use 4 feet (1.2m) of ordinary shielded audio cable and find that 8 turns around a ¼ inch (6.5mm) polystyrene coil form works well up to the frequency limit of the generator (40 MHz). An RCA phono plug is mounted on the coil form.

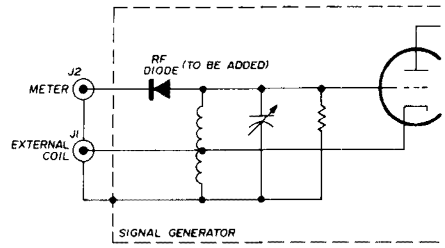


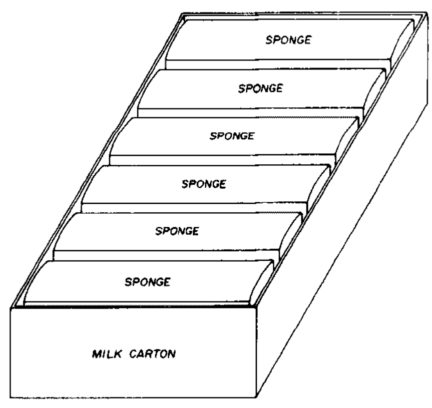
fig. 4. The rf pickup coil is rectified by the diode and indicated by the meter.

When you are ready to test the absorption device, couple its search coil to a grid-dip meter. At resonance, the meter will read about full-scale with the coils ½ inch (13mm) apart. Turn off the signal generator when making absorption measurements.

I. Queen, W2OUX

soldering tip cleaner

A clean soldering-iron tip helps make reliable solder connections. Fig. 5 shows an inexpensive tip cleaner made from a half-pint (¼ liter) milk container and some sponges cut to fit the container.



SOLDERING IRON TIP CLEANER

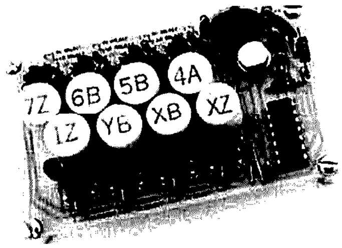
fig. 5. Soldering iron tip cleaner.

Best results are obtained when the sponges are kept moist. To clean your soldering iron, just insert the tip between the sponges.

Gary Tater, W3HUC

NEW products

sub-audible tone encoder



Communications Specialists announce their new ME-8, a sub-audible tone encoder for repeater access. Measuring only 2x3x9 inches (5x7.5x22.5cm) and operating on any voltage between 6 and 16 volts dc, the unit is capable of generating any eight of the 32 EIA tone codes between 67.0 and 203.5 Hz.

Frequency selection is made by supplying ground or positive voltage to the desired control lead, and plugging in the desired K-1 element. The unit is completely immune to rf and has a start-up time of only 10 milliseconds. Frequency accuracy and stability are excellent and the output is a low-distortion sinewave at 3 volts rms.

The ME-8 is priced at $79.95 and includes eight K-1 elements. Extra elements are available for $3.00 each. For additional information write Communications Specialists, Post Office Box 153, La Brea, California 92621 or use *check-off* on page 134.

high-current regulated power supplies

VHF Engineering, Binghamton, New York, has announced two new regulated 12-volt dc power supplies for use in amateur and commercial applications. The two power supplies have current ratings of 15 amps at 10 to 12 volts dc, and 25 amps at 10 to 12 volts dc, respectively. Voltage regulation is 2% from no load to maximum rated current.

These new power supplies feature over-voltage protection and fold-back current limiting. Over-voltage protection is provided by a crowbar circuit which will shut down the supply if its output voltage exceeds 14 volts, thereby protecting the equipment being powered. The foldback current-limiting circuit limits the output current to a maximum of 1 amp if the output is shorted. This protects the power supply from damage and prevents high current from flowing into a piece of equipment that may have developed an internal short. When the short is removed, the power supply output returns to normal.

These regulated power supplies have been designed to commercial standards and use tinned, glass-filled epoxy circuit boards and high quality components. Additional information may be obtained from VHF Engineering, 320 Water Street, P.O. Box 1921, Binghamton, New York 13902 or use *check-off* on page 134.

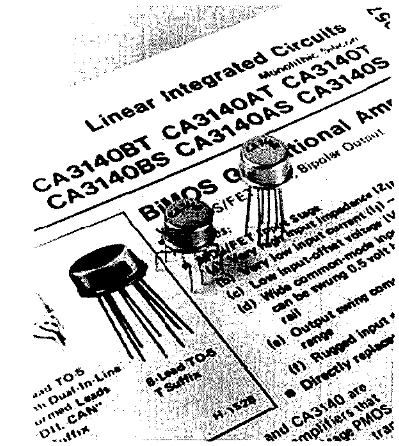
air-variable capacitors

Among the components which are most difficult to find these days are air-variable capacitors, especially those rated for transmitting powers. Now, however, the Dentron Radio Company is offering a line of air variables which are suitable for most amateur applications in transmitters and antenna tuning units. Listed in the table above are the electrical specifications for these capacitors.

model	type	capacitance minimum	capacitance maximum	voltage rating	price
D-88-120	single	53 pF	208 pF	6000	$23.50
D-140-75	single	23 pF	140 pF	4500	16.50
D-232-45	single	23 pF	232 pF	3000	16.50
D-500-45	single	48 pF	500 pF	3000	19.50
DD-150	dual	33 pF (per section)	205 pF	3000	20.00

These air-variable capacitors may be ordered directly from Dentron Radio Company, 2100 Enterprise Parkway, Twinsburg, Ohio 44087. For more information use *check-off* on page 134.

bipolar/mos op amp



A new bipolar/mos op amp, the CA3140, featuring a pmos input stage and a bipolar output stage with a wide output voltage range, has been introduced by RCA Solid State Division. Its versatility permits it to fill virtually all 741 sockets and, at the same time, most of the premium op amp sockets currently served. The pmos input stage is similar to the one used in the RCA CA3130 op amp but with internal compensation and the ability to operate from a supply voltage between 4 and 44 volts, dual or single supply. A special feature is the addition of bipolar diodes which protect the input to such an extent that preliminary tests, under simulated electrostatic conditions up to 1000 volts, show the CA3140 to be more rugged than any other device yet tested, including bipolar and fet input op amps.

Typical performance features for the CA3140 include: an input impedance of 1500 megohms; input current of 10 pA at ± 15 volts, low input-offset voltage of

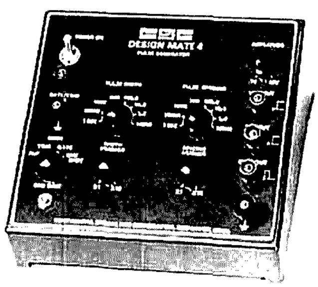
More Details? CHECK-OFF Page 134

5 mV, wide common-mode input voltage range, -0.5V below negative bus; output swing to within 0.2 volt of negative supply; high and low operating supply voltage of 4 to 44 volts, high slew rate of 9V/μs; high gain bandwidth product of 4.5 MHz; and fast settling time of 1.4 μs, to 10 mV with a 10 volt$_{p\text{-}p}$ signal. The new device is available in the basic TO-5 package or the dual-in-line package.

The breadth of applications permits the device to replace numerous op amp categories such as general purpose, fet-input, and wideband or high slew-rate types. The wide bandwidth feature reduces costs by permitting wideband video and audio circuits at lower cost than with current op amps, as well as lower-cost wideband TTL interfaces. The strobable output stage allows the output to be driven low, independent of the input signal. The fact that the output swings to within 0.2 volt of the negative supply permits power transistors to be driven directly, thus eliminating level shifting circuitry.

For additional information, including copies of the 20-page data sheet, File no. 957, and a descriptive applications brochure, Publication no. 2M1144, offering a free sample, write to RCA Solid State Division, Box 3200, Somerville, New Jersey 08876 or use *check-off* on page 134.

digital pulse generator



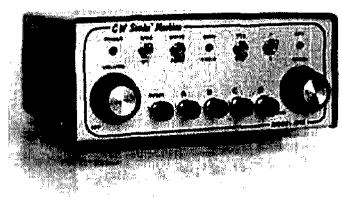
Continental Specialties Corporation has introduced the Design Mate-4, a versatile, laboratory-quality generator which lists for $124.95. The DM-4 may be used as a clock source, delayed pulse generator, synchronous clock source, manual system stepper, pulse stretcher, or clock burst generator.

The DM-4 is designed to fill the bill wherever a source of clean, crisp, fast output pulses compatible with virtually

all logic families and descrete circuits is required. It is capable of generating symmetrical and asymmetrical pulses from 0.5 Hz - 5 MHz and has a positive output of 100 mV to 10 V, with a rise and fall time of less than 30 nanoseconds. Additionally, the DM-4 offers an independently controlled pulse width and spacing from 100 ns to 1 second in seven overlapping ranges, as well as independent variable-amplitude CMOS, and fixed-amplitude TTL outputs. The unit operates either continuously or in manual one-shot fashion. It also features external triggering from dc to 10 MHz and synchronous output gating.

For more information contact Continental Specialties Corporation, 44 Kendall Street, Box 1942, New Haven, Connecticut 06509 or use *check-off* on page 134.

electronic keyer



The *CW Sendin' Machine* was designed by an amateur and allows you to write in your log while it sends your routine information. Not only a fine keyer with iambic operation and dot and dash memories, it is also equipped with two random-access memories that can store short CW messages and send them at the push of a button.

The model 2048 is equipped with a cord and plug to be connected to your favorite paddle, The keyer contains no relays and can operate into open-key voltages of up to 150 volts and closed-key currents to 10 mA. Sidetone level can be controlled by a volume control; pitch can be varied by a pot on the PC board.

The *CW Sendin' Machine* model 2048 comes with two 1024-bit memories, and an automatic reset feature is available as a modification which can be added to your board. The PC board is on a connector for easy service. The board operates on 5 volts at 400 mA.

For more information write H.A. Harp, WA4SVH, 718 Magnolia Dr., Lake Park, Florida 33403 or use *check-off* on page 134.

More Details? CHECK-OFF Page 134

short circuits

frequency synthesizers

Several errors crept into DJ2LR's excellent article on frequency synthesizers in the July, 1976, issue of *ham radio*. In **fig.** 13 (page 16), the wiper of the 4.7k potentiometer between the emitters of the two 2N3570 transistors should be grounded. In **fig. 17** the input labeled "10 kHz from divider" should show a pulse width of 200-500 nanoseconds (*not* microseconds). In **fig.** 20 the CD4000 gate should be a 74COO, and in **fig. 23** the gate with three inputs which is not identified is a 74LS30. Finally, the caption to fig. 24 on page 23 is incorrect – this circuit *adds* 41 MHz to the synthesizer reading.

DT-600 RTTY demodulator

In the schematic of the DT-600 RTTY demodulator in the February, 1976, issue of *ham radio* one end of the 10k balance pot (at the input to U1, wiper connected to R9) should go to -12 Vdc. This pot should be labeled R8. In **fig.** 5 the horizontal axis was measured in Hz but mistakenly labeled in baud; none of the standard RTTY speeds (60, 75, and 100 wpm) are attenuated. Potentiometers to fit the circuit board are available from Data Technology Associates, Inc., Box 431912, Miami, Florida 33143. A set of four pots is priced at $2.00.

DT-500 RTTY demodulator

Due to an editorial oversight, the names of co-authors Garey K. Barrell, K4OAH, and Archie C. Lamb, WB4KUR, were inadvertently deleted from the cover page of the DT-500 article in the March, 1976, issue. In addition, in **fig. 2** (page 26), the negative return from the +170 volt loop supply should be connected to ground through a 2500 ohm, 20 watt resistor. Printed-circuit boards for the DT-500 are priced at $10.50 from Data Technology Associates, Inc., Box 431912, Miami, Florida 33143. A set of four PC potentiometers is available from the same source for $2.00.

S-line frequency synthesizer

In the S-line frequency synthesizer article published in the December, 1975, issue, in **fig.** 3 (page 12) R2 should be connected to pins 4 and 5 of U1B, *not* pin 6. The parts layout in **fig.** 10 is correct.

More Details? CHECK-OFF Page 134

one dollar

ham radio magazine

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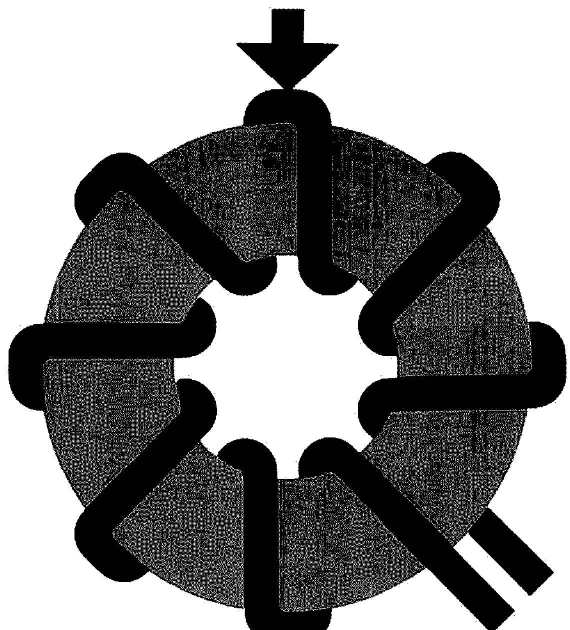
wideband rf autotransformers



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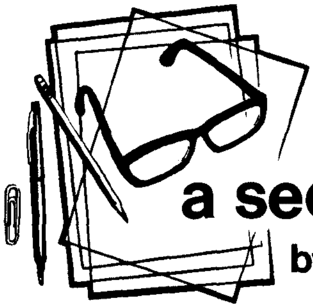

a second look

by Jim Fisk

Interference on the amateur bands is something that most of us have learned to live with, at least to a certain extent, but in recent months I have noticed an increasing number of bad operating practices cropping up on our bands. Apparently other amateurs have been troubled, too, because I have received a number of letters on the subject. None of these practices is new, but they're more offensive because the bands are much more crowded than they used to be. Deliberate interference, tuning up on net frequencies, playing music, calling CQ without listening first, offensive language, incorrect identification (or no identification at all), using a kilowatt when 100 watts is adequate, talking crosstown on 20 meters instead of using vhf-fm – the list could go on and on.

There's no question that our high-frequency bands are crowded, but deliberate and malicious interference, and discourteous operating tactics, aren't going to relieve the situation, anymore than elbowing your way to the front of the checkout line at the local supermarket is going to get you anything more than a fat lip! In case you hadn't noticed, everything is more crowded today – the population has exploded, the expressways and turnpikes are jammed, homes are being built on smaller and smaller pieces of land, and even on the remotest trail of the high Sierras it's impossible to escape from the inevitable discarded beer can – practically everywhere you go you find a mass of humanity. It follows that we'll have more and more congestion on the amateur bands, but congestion doesn't mean bedlam. Zeroing your kilowatt on a QSO or local net is not going to make them move. Why not join them? They'd probably be glad to have you.

Today, there is a net for almost every range of interest – they aren't restricted to handling traffic. Some of the groups that congregate on the bands are not really nets at all, but simply groups of hams who get together for a common purpose. There are DX nets, the county hunters, the early-morning groups on 75 meters, and various single-frequency gabfests. There are technical nets, satellite operators' nets, EME round tables, vhf nets, and, of course, a multitude of local and intercontinental traffic nets, if traffic handling happens to be your *forte*.

If you don't happen to care for net-type operation, fine; there are a good many amateurs who don't. On the other hand, if there weren't any nets, imagine what the QRM would be like. There are thousands of amateurs who congregate on particular net frequencies. Since they're a member of a net, they just "read the mail" a good deal of the time. If they didn't have the net, they would be calling CQ, fishing for a new county, or active in one of the horrendous DX pileups. So, when you hear a net in operation, don't use it for a tuneup frequency. Whether you know it or not, the most hedonistic of them will stand to handle emergency traffic if asked to do so. They *all* do a service to the amateur community by minimizing interference with channelized communications.

Deliberate interference and incorrect identification are only two of the bad operating practices you will find on any band you listen to. You can also hear any number of stations working cross town on 15 or 20 meters when they should be on 75 or vhf fm. I have often copied W/K stations on 20 meters, running well over S9 in New Hampshire, working their neighbors. With modern linear amplifiers, it's a simple matter to turn off the big box when you don't need it.

Why the big penchant for S9 signal reports, anyway, when you can maintain perfectly adequate QSOs with S6 or S7? You may need your linear for a long-haul DX QSO or for making initial contact, but once communication has been established, 95 percent of the time you can turn your linear off with absolutely no effect on the QSO. Owning a linear is a bit like carrying an umbrella to work every day – there are times when it's a practical necessity, but just because you own one doesn't mean you have to use it all the time.

I've heard a lot of stations go QRT because of interference and poor operating practices. This is not the answer. If you hear an amateur on the air with a bad signal, not identifying properly, causing unnecessary interference, or being generally obnoxious, tactfully tell him about it. Most amateurs are gentlemen and will accept your suggestions with grace.

And, when you go on the air the next time, use operating finesse instead of brute force. Strive to be a first-class operator and encourage your friends to do the same. Let's promote good operating on our bands – discourtesy breeds pandemonium.

Jim Fisk, W1DTY
editor-in-chief

presstop

"HAM RADIO HORIZONS," a new Amateur Radio magazine aimed primarily at the newcomer, has been announced by Ham Radio and will be introduced in January. Editorially the new publication will be directed toward the beginning Amateur, with low level technical articles plus general interest features for Amateurs and SWL enthusiasts at all levels. As an attraction to potential Amateurs, some CB and SWL coverage will also be included with the expectation that readers seriously interested in those fields are very good prospects for conversion to Amateur Radio.

Very Broad Market penetration will be a feature of Ham Radio's new publication. Promotion in a variety of non-Amateur Radio media is planned and emphasis will be on wide distribution far beyond the usual radio store outlets. Since it will be written for and by Amateurs, its content will appeal to most Amateurs, old timers as well as Novices, but its primary goal is to attract new Amateurs and encourage their progress up through the Amateur ranks.

Subscriptions To Ham Radio Horizons are now being accepted at $7.00 per year; after January 1st the cost for a one-year subscription will increase to $10.00.

Ham Radio Horizons Already Has a good backlog of technical articles but it is looking for good fiction, humor, and other non-technical contributions. Prospective authors should contact Tom McMullen, W1SL, Managing Editor, Ham Radio Horizons, Greenville, New Hampshire 03048.

FIRST U.S.-BERMUDA TWO METER CONTACT was made by W1NU/VP9 and K1HTV September 14th at 0100Z! Alerted to a potential opening by the AMSAT Net, Vic and hopeful stateside two-meter DXers moved to 144.090 CW and within moments the historic contact began. Vic next worked K1WHS, and before the evening ended 30 more stations, from Maine to Pennsylvania, at least one running only 100 watts to an 11-element beam, had made it across the Atlantic to Bermuda.

FCC LICENSE TIMES continue to improve, with some Amateurs receiving theirs in as little as six weeks while most CB tickets are showing up in just over half that time. Applicants whose licenses don't arrive in a reasonable time might check with their banks — John Johnston said in Hawaii that FCC is getting about 300 rubber checks each week, but very few from Amateurs!

AN UNLICENSED CW OPERATOR can operate an Amateur station under a current interpretation of the rules. However, it must be with a control operator present and the unlicensed operator's transmissions logged and recorded as third party traffic.

A TVI-LESS TELEVISION SET is to be built by Texas Instruments under contract from the FCC, according to FCC Chief Engineer Ray Spence. The experimental set will provide graphic demonstration that TVI is not all the fault of the transmitter. Consumer complaints have shifted recently, however, and audio component systems have now become more of a problem than TV receivers.

PLANNING FOR WARC 79 is still very much in the news. Amateur Radio, as defined by Article 41, will not be discussed at the 1979 Geneva Conference — the proposed agenda includes only Article 5, the international frequency table. We will face problems — our own broadcasters want parts of 40, 75, and 160, and now the Canadian DOT wants to make 1605-4000 kHz all maritime and the CBC wants to shove 40 meters down to 6800-7100 kHz so 7100 and up can be devoted to international broadcast!

Clearly More Supportive Work needs to be done, both with our own government and with other governments friendly to the Amateur Radio Service.

THE "PERSONAL COMMUNICATIONS RESEARCH ASSOCIATION" was formed in Los Angeles in early September by ten legal people who met in Jon Gallo's (WA6PTM) law offices. The non-profit organization, eventually to become the "Personal Communications Research Institute," will be a coordinating body for attorneys who find themselves in legal conflicts involving the personal communications field, primarily Amateur Radio and Citizens Band.

Tower Restrictions, anti-TVI/RFI ordinances, and dealings with various governmental agencies will be the foundation's prime concerns. If funding permits, it would also sponsor major research studies supporting the concept of personal communications.

Long Term Support will be sought from both industry and individuals but start-up costs are being borne by the organizers. All are Amateurs (half with CB licenses) with experience in helping support the concept that Amateur and CB radio is a "reasonable and proper" use of an individual's home.

10,000 MHZ DX RECORD was broken by two U.K. Amateurs in August when G4BRS in Cornwall worked GM3OXX in Edinburgh, a distance of 323 miles (521 km). Despite the much larger U.S. Amateur population, the U.K. and other European countries seem to be much more active on 3 cm and other Amateur microwave bands.

wideband rf autotransformers

Design data for broadband impedance transformers using a single core

The use of transmission-line transformers wound on toroidal forms has become very popular in the last few years. These transformers serve many useful purposes, probably the most common being the transformation of balanced to unbalanced circuits and vice versa (baluns) over a relatively wide bandwidth. They were first described by G. Guanella.[1] The basic paper on the subject in this country was written by Ruthroff.[2]

Transmission-line transformers are restricted, however, as to available impedance transformation ratios and circuit configurations. Some of these restrictions can be relaxed by using an autotransformer instead of a transmission-line transformer, for although the construction techniques for the two types of transformers are similar, there are some very subtle differences in their operation, design, and application.

The purposes of this article are to point out the differences between transmission-line transformers and the more conventional autotransformers, and to give some design considerations for autotransformers. Autotransformers have many applications that *can't be* realized with transmission-line transformers, such as unbalanced-to-unbalanced or balanced-to-balanced impedance transformations, or impedance transformation ratios other than an integer squared. Some impedance transformation ratios that can be obtained with transmission-line transformers by using multiple cores can be obtained with a single core by using an autotransformer.

transmission-line transformers

The distinguishing feature of the transmission-line transformer is that the "winding" is composed of two conductors with equal and opposite currents flowing in each conductor, as in a balanced transmission line. The net magnetizing ampere-turns in the core is zero. Because of this, there is considerable difference in the means by which energy is transformed between the input and output circuits for a transmission-line transformer and an autotransformer.

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Quoting from reference 3 in describing a transmission-line transformer, "the inductance of a conductor is directly proportional to the relative permeability of the surrounding medium. A high permeability material placed close to the transmission line conductors acts on the external fringe field present and can magnify the inductance appreciably and thereby provide lower cutoff frequency. There is no influence upon internal magnetic fields nor upon the characteristic impedance of the line. The power transferred from input to output is not coupled through the ferrite material, but rather through the dielectric medium separating the transmission line conductors. This is an important concept for design. Relatively small cross-section ferrite material can perform unsaturated at impressively high power levels. In contrast to this, conventional transformers couple power from primary to secondary entirely through a high permeability core which must be chosen to suitably carry the total power without saturating."

With the autotransformer, the permeability of the core and the number of turns determine the low-frequency response, while the high-frequency response of these transformers is obtained by very close coupling, both capacitively and magnetically, between the windings. This is obtained by the twist.

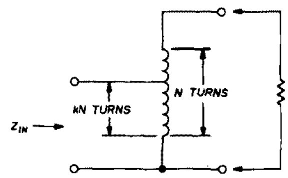


fig. 1. Basic autotransformer circuit used as a step-down device. In this application $Z_{in} = k^2Z_o$ for k between zero and 1, where Z_{in} and Z_o are input and output (load) impedances and k is the tap location, expressed as a fraction of the total number of turns referenced to the common terminal.

autotransformers

The basic autotransformer circuit is given in **fig. 1.** When the transformer is used as an impedance step-down device, the relationship between the load impedance and that presented to the input terminals is given by

$$Z_{in} = k^2 Z_o \quad (1)$$

where Z_o is the load or output impedance
Z_{in} is the impedance presented to the input terminals
k is the location of the tap (stated as a fraction of the total number of turns referenced to the common terminal).

By John J. Nagle, K4KJ, 12330 Lawyers Road, Herndon, Virginia 22070

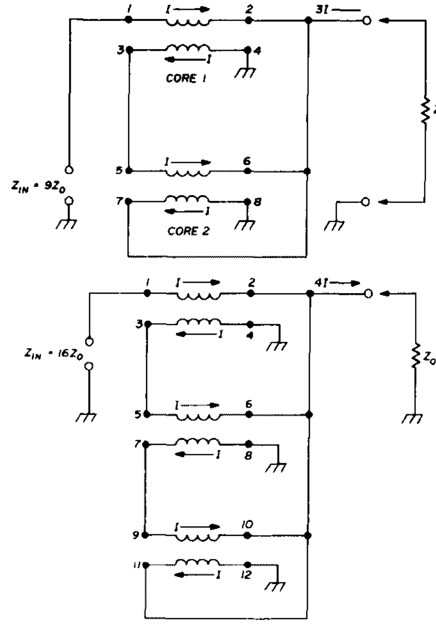


fig. 2. Transmission-line transformers for unbalanced-to-unbalanced loads. A and B are for 9:1 and 16:1 transformation ratios respectively. (Data is from reference 3).

When using transmission-line transformers, it's necessary to have two identical windings. The tap must be placed at the center, giving a four-to-one impedance transformation. This case has been well covered in the literature and will not be considered further. Unfortunately, impedance ratios other than four-to-one are frequently required, and methods of obtaining both integral and nonintegral impedance ratios are not so well known.

unbalanced-to-unbalanced loads

Pitzalis and Couse[3] give circuits for obtaining 9:1 and 16:1 impedance ratios for unbalanced-to-unbalanced loads. These circuits are shown in figs. **2A** and **2B** respectively. The 9:1 transformer requires two cores, while the 16:1 unit requires three cores; in these circuits, there is one bifilar winding on each core.

Because of the requirement for equal and opposite currents in the windings, transmission-line transformers are restricted to impedance ratios of $(\frac{2}{1})^2, (\frac{3}{1})^2, (\frac{4}{1})^2 \ldots (\frac{n}{1})^2$ where n is an integer. In general, the impedance transformations available with this type of circuit are

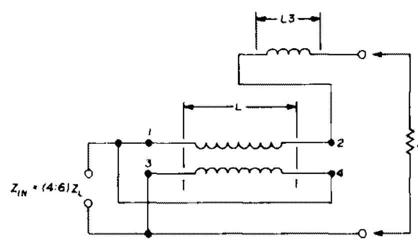


fig. 3. Method for extending the impedance-transformation ratio of an unbalanced-to-unbalanced transformer from 4:1 to about 6:1 (after Krause and Allen, reference 4).

given by

$$impedance\ ratio = (N_c + 1)^2 \qquad \textbf{(2)}$$

where N_c is the number of cores

Impedance transformation ratios other than those obtainable from transmission-line transformers are frequently required, and so the autotransformer must be considered. Krause and Allen[4] have given an interesting method for extending a 4:1 transformation ratio to 6:1. After the usual bifilar winding is wound on the core, a third winding is wound over the original winding, as shown in fig. **3**. The length of this winding is given by

$$L3 = L\left[\sqrt{R_L/R_S} - 2\right] \qquad \textbf{(3)}$$

where $L3$ is the length of the third winding
R_L is the load resistance
L is the length of the bifilar winding
R_S is the source resistance

If impedance transformations greater than about 6:1 are attempted, bandwidth reductions may occur.

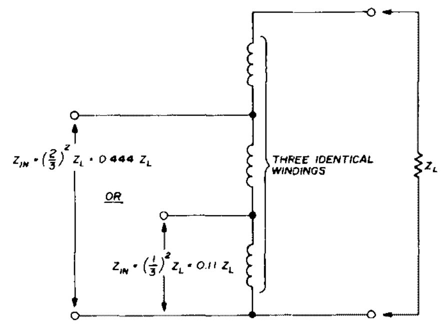


fig. 4. Autotransformer using a trifilar winding gives impedance step-up ratios of 1:9/4 or 1:9, depending on which tap is used.

The addition of this third winding changes the device from a transmission-line transformer to an autotransformer, since the tertiary winding has no conductor to carry an equal and opposite current. Care must be taken to ensure the core does not become saturated by the unbalanced current flow.

The characteristic of transmission-line transformers cited previously is also evident from fig. 2; in a true transmission-line transformer, the current flow must be equal and opposite in the two conductors on the core, just as in a transmission line. Many of the devices one sees parading as transmission-line transformers, including some to be described later, are not true transmission-line devices but are simply broadband transformers of the autotransformer variety.

It's possible to obtain transformation ratios of other than 4:1 by autotransformer techniques; these have the further advantage that only one core is needed. If a trifilar winding is made by tightly twisting three separate conductors together to form a single winding and the three conductors connected in series aiding, the circuit of fig. 4 results. This gives impedance step-up ratios of $1{:}\frac{9}{4}$ or 1:9, depending on which tap is used. This device

is of the autotransformer type and is not a transmission-line transformer since the currents are not equal and opposite through the individual conductors, although the construction techniques are similar and the bandwidths obtainable appear to approach those of a transmission-line transformer.

Going one step further and using a quadrifilar winding, impedance step-up ratios of $1{:}(3/4)^2$, $1{:}(1/2)^2$, and $1{:}(1/4)^2$ can be obtained, as shown in fig. 5, thereby increasing the flexibility afforded the circuit designer. It's difficult to say how much further the process can be continued, but the possibilities appear limited by the number of wires it's practical to put on a core. It should also be expected that the usable bandwidth will decrease as the number of windings increases.

The type of construction just described has the

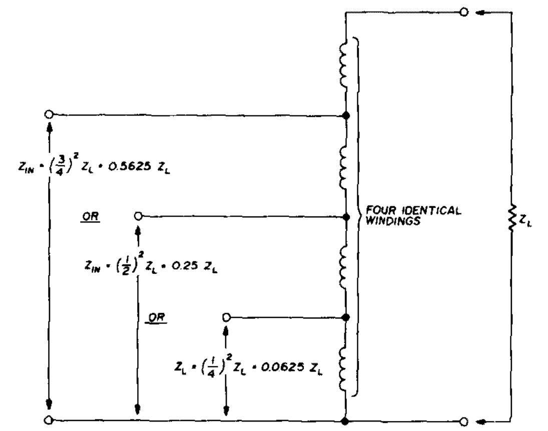


fig. 5. Autotransformer with a quadrifilar winding will give step-up ratios of $1{:}(3/4)^2$, $1{:}(1/2)^2$, and $1{:}(1/4)^2$.

advantages of accuracy and convenience. As broadband rf transformers of the type described have only a very small number of turns, it's difficult to place a tap accurately. With the transformers to be described, the available transformation ratios depend only on the number of windings put on a core, so that miscounting a turn or two won't cause an error. The total number of turns can be selected to obtain the required inductance. The arrangement is convenient because the taps are placed at the junction of two windings, which is easily accessible. A further advantage is that the impedance transformation ratios available are the square of a rational number instead of the square of an integer, as with transmission-line transformers.

balanced-to-balanced loads

In the preceding material we've assumed the unbalanced-to-unbalanced case. When three or more windings are put on the same core, as in figs. **6** or **7**, the transformer may also be used for balanced-to-balanced impedance transformations. A three-winding core is shown in fig. **6** which will give an impedance ratio of 0.111:1; a four-winding core will give an impedance ratio of 0.25:1; see fig. 7. An advantage of using an even number of windings is that the center tap may be grounded.

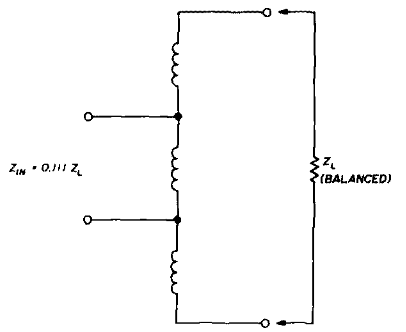


fig. 6. A trifilar-wound transformer that provides a balanced-to-balanced impedance transformation of 0.111:1.

It's desirable to be able to determine early in the design stage whether a given impedance transformation can be obtained with an n-filar winding with taps at the junctions or whether a tapped winding must be used. The following slide-rule algorithm will easily determine this:

Assume it's desired to match a 50-ohm coaxial cable to a 72-ohm load using an autotransformer. Set 50 on the *A* scale of a slide rule opposite 72 on the *B* scale. Now scan the *C* and *D* scales for two integers that line up; notice that **6** on the *C* scale is opposite 5 on the *D* scale (fig. **8**). This indicates that a hexifilar winding connected across the 72-ohm load, with the 50-ohm cable connected between the fifth and sixth winding, will be required as shown in **fig. 9**. A resistive pad used to obtain the same impedance match would require a loss of 5.7 dB.

The same procedure may be used for balanced lines except that in scanning the *C* and *D* scales of the slide rule, it's necessary to look for integers separated by two or any other even number. For example, let's say we wish to match a 383-ohm balanced line of AWG 18 (1mm) wire spaced ½ inch (12.5mm), such as open-wire TV line, to 200 ohms balanced (200 ohms was chosen since it's the impedance obtained in going from 50-ohm coax through a 4:1 balun).

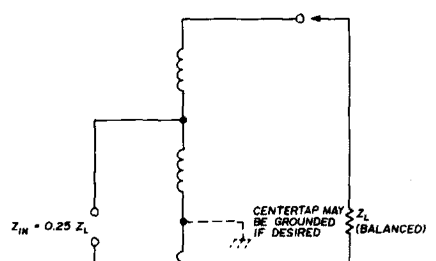


fig. 7. A balanced-to-balanced impedance transformation of 0.25:1 may be obtained with a quadrifilar winding. The center tap may also be grounded.

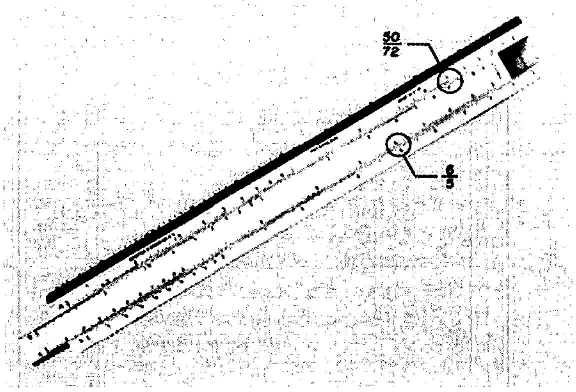


fig. 8. Slide-rule algorithm for determining the turns ratio for matching 50 ohms to a 72-ohm load. Impedance ratios on the A and B scales show that the turns ratio should be 5:6 (C and D scales).

After setting 200 on the *B* scale opposite 383 on the *A* scale, we note that the closest integers separated by two are 7 on the *D* scale opposite 5.05 on the *C* scale (fig. 10). Because only integral numbers are allowed, an exact match can't be obtained. It is therefore necessary to move the slide slightly so that 5 and 7 are aligned on the *C* and *D* scales. This puts 200 on the *B* scale opposite 390 on the *A* scale, and will give a vswr of 390/383 = 1.02:1 (this may or may not be acceptable, depending on the application). At any rate, it's the best that can be done with an n-filar winding. If a better match is required, a tapped winding must be used. The circuit is shown in fig. 11.

A note of caution is necessary in applying this algorithm. Make certain that the proper side of the *A* and *B* scales is used in setting up the impedance ratio. In the first example above, both impedances were in the decade between 10 and 100, while in the second example both impedances were in the decade between 100 and 1000, so that in both cases the same half of the *A* and *B* scales would be used. If, on the other hand, you re trying to match 75 to 120 ohms (both unbalanced), you should use 75 from the left-hand side of the *A* scale with 120 from the right-hand of the *B* scale and obtain 5.05 over 4 on the *C* and *D* scales. If you mistakenly set 75 on the left-hand *A* scale opposite 120 on the left-hand *B* scale, the result would be 2 over 5 on the *C* and *D* scales, which is incorrect. This is shown in fig. 12.

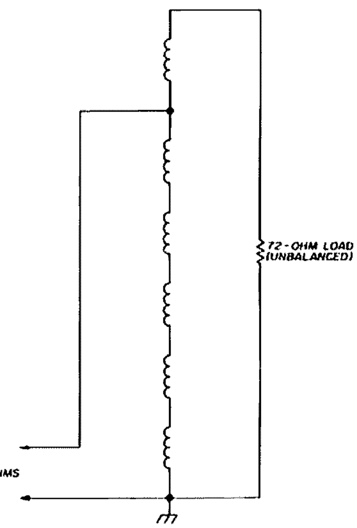


fig. 9. Hexifilar winding on an autotransformer for matching a 72-ohm load to a 50-ohm transmission line. Circuit illustrates the tap placement in the example of fig. 8.

I believe the above algorithm using a slide rule is more convenient to use than trying to calculate an exact turns ratio on a pocket calculator, even a programmable type. A slide rule can tell you at a glance the possible combinations that will give the desired turns ratio and how close it's possible to come if there are no exact ratios.

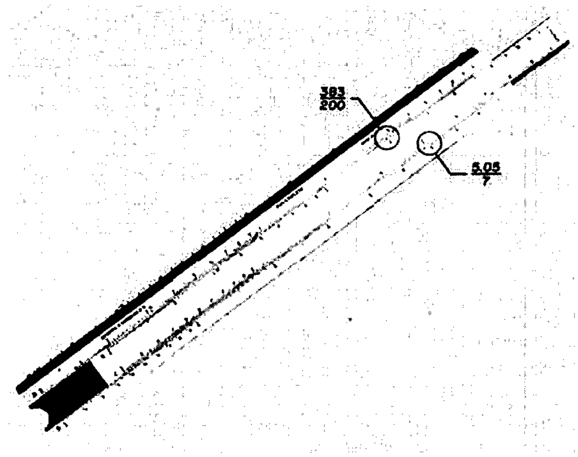


fig. 10. Slide-rule setup shows a turns ratio of 5:7 will match 200 to 383 ohms (approximately), both balanced to ground.

Also the more common four-function calculators won't square nor extract square roots, which you'll need to go from turns ratio to impedance ratios, or vice versa.

core configuration

One of the important parameters in the design of a successful wideband rf transformer is the core shape. Most neophytes in transformer design, including me, begin by using toroids, probably from tradition. Toroids, however, are seldom the best choice. We will, therefore, discuss the subject of core shape briefly; the following material was taken from references 5 and 6.

The factors limiting the high-frequency response of a transformer are leakage inductance and winding capacitance; these factors suggest a winding that has as few turns as possible. On the other hand, the low-frequency response is limited by the available shunt inductance, and this suggests that a large number of turns may be necessary to meet the low-frequency specifications. It is therefore of considerable practical interest to find a core shape that will maximize the shunt inductance while minimizing the leakage inductance and shunt capacity, and to rate various core shapes numerically on this basis.

Such a number is called the form factor and is defined as

$$form\ factor = \frac{l_w l_e}{A_e} \qquad (4)$$

where l_w is length of one complete turn of wire
l_e is effective length of the magnetic circuit
A_e is effective cross sectional area of the core

All the above factors must be in the same units, usually millimeters. The smaller the form-factor number, the more desirable the core for high frequency, broadband transformer use.

In most core manufacturers' literature, the quotient l_e/A_e is given its own symbol, usually C_I, so that

$$form\ factor = l_w C_I \qquad (5)$$

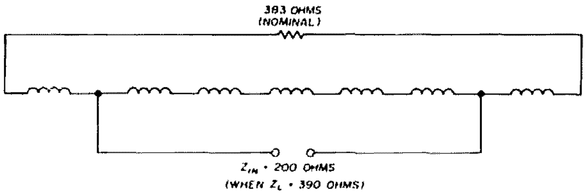


fig. 11. A 7-winding autotransformer for matching a 200- to 383-ohm load.

It's of interest to consider briefly the form factor for a toroid and various ways to minimize it. This will lead to a more optimum core shape. A detailed discussion is given in reference 6, pages 265-267.

The length of a single turn of wire, l_w, for a toroid is given by

$$l_w = d_2 - d_1 + 2h \qquad (6)$$

where d_2 and d_1 are the outer and inner diameters of the core respectively

h is the axial length (height)

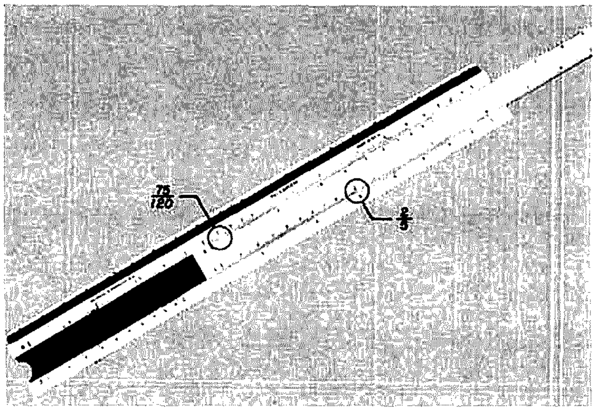


fig. 12. You must watch the slide-rule scale settings carefully or errors will result. This example shows the slide rule set to the wrong half of the A or B scales, which gives an incorrect turns ratio for matching 75 to 120 ohms.

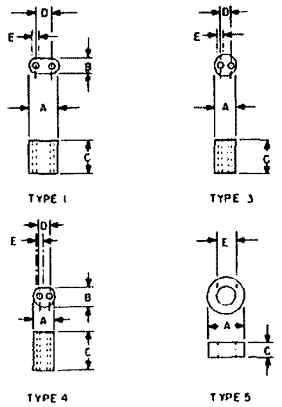
The factor C_I is given by Snelling[6] as

$$C_I = \frac{2\pi}{h \log e \frac{d_2}{d_1}} \qquad (7)$$

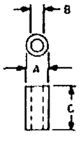
The objective is to minimize the product $l_w C_I$; this can be done by minimizing either or both factors. C_I can be decreased by either increasing d_2 or decreasing d_1. However, when the ratio $\frac{d_2}{d_1} \gg 1$, increasing d_2 increases l_w far more than a proportional decrease in d_1 so that it's desirable to make d_1 as small as possible. As only a few turns of wire are used for most broadband rf transformers, the inner diameter can usually be reduced to a very few millimeters. This results in a toroidal form quite different in shape from the usual toroid and is more often referred to as a bead.

There is a practical limit, however, as to how far the form factor may be reduced in this manner. Further improvement can be made by placing two toroids edge to edge, as shown in fig. 13, and threading turns through

table 1. Form factors for commonly used toroid and balun core sizes.



core shape type	A	B	C	D	E	form factor
1	0.525(13.3)	0.295(7.5)	0.407(10.3)	0.225(5.7)	0.150(3.8)	13.0
1	0.277(7.0)	0.160(4.1)	0.244(6.2)	0.114(3.6)	0.071(1.8)	14.3
1	0.136(3.5)	0.079(2.0)	0.093(2.4)	0.057(1.4)	0.034(0.9)	14.0
3	0.250(6.4)	—	0.242(6.1)	0.100(2.5)	0.050(1.3)	9.5
3	0.250(6.4)	—	0.471(12.0)	0.100(2.5)	0.050(1.3)	8.8
4	0.284(7.2)	—	0.218(5.5)	0.104(2.6)	0.052(1.3)	8.8
3	0.220(5.6)	—	0.250(6.4)	0.090(2.3)	0.035(0.9)	7.8
5	0.380(9.7)	—	0.190(4.8)	—	0.197(5.0)	29.0



nominal dimensions, in. (mm) A	B	C	form factor
0.138(3.5)	0.051(1.3)	0.118(3.0)	17.3
0.076(1.9)	0.043(1.1)	0.150(3.8)	24.3
0.138(3.5)	0.051(1.3)	0.236(6.0)	14.9
0.138(3.5)	0.051(1.3)	0.500(12.7)	13.7
0.296(7.5)	0.094(2.4)	0.297(7.5)	14.8
0.200(5.0)	0.062(1.6)	0.250(6.4)	15.2
0.200(5.0)	0.062(1.6)	0.437(11.0)	12.3

the two holes. In practice, a single core would be used having the shape shown in fig. 14. This type of core is often referred to as a balun core. A disadvantage of this core shape is that each turn must be threaded through two holes, but since the number of turns is small, this isn't a serious limitation. Table 1 gives the form factor for various commonly used toroid and balun core sizes. This data was taken from reference 5. From this table, it can be seen that the balun shape has a lower form factor than the toroid.

low-frequency response

The design of wideband autotransformers is similar to that of transmission-line transformers. Although both are relatively straightforward, some compromises must be made from a practical standpoint. For the autotransformer, the desired low-frequency response determines the number of turns and the size of the core. The low-frequency response will be down 3 dB at a frequency, f_1, when the primary reactance is equal to the total shunt resistance. The shunt resistance is the parallel combination of a) the load impedance referred to the primary, b) the parallel equivalent of the generator resistance, and c) the transformer losses expressed as a shunt resistance across the primary. These are shown in fig. 15. This will occur when $2\pi f_1 L_p = R_{eq}$ where R_{eq} is the parallel combination of the load, generator and loss resistances referred to the primary. Solving for L_p gives

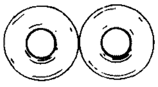


fig. 13. Placing two toroid cores side by side gives a better form factor than a single toroid. The winding is through the two holes.

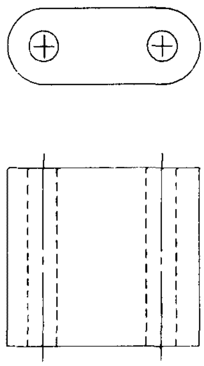
$$L_p = \frac{k^2 R_{eq}}{2\pi f_1} \qquad (8)$$

The response will be down 1 dB at a frequency $3f_1$. Most core manufacturers publish a factor, A_L, which gives the inductance per turn (or per turns squared) or per hundred turns. In the megahertz region, A_L varies with frequency so this factor is usually presented as a curve. A_L should be chosen for the lowest frequency of interest. Using this factor and the inductance found from eq. 8, the required number of turns can be calculated. If it's not possible to wind the desired number of turns on the selected core, it will be necessary to use a larger core or a core with a higher permeability, thereby requiring fewer turns. Where wide temperature variations are possible, cores with high permeability should be avoided, as these cores also have a higher temperature coefficient than cores with lower permeability, which may cause problems at the temperature extremes.

high-frequency response

The high-frequency response is a different problem. In general, to maintain good high-frequency response, the number of turns should be as small as possible. Thus some juggling is necessary among the number of turns, core permeability, and core size and shape.

fig. 14. A balun core as developed from two toroidal cores placed side by side.



design example

Transformers of this type are not difficult to wind. If you've never tried winding one before, you may have to wind two or three before you get one you like, but keep trying; analyze your deficiencies in light of the above material. When you find the right combination, you'll probably be pleasantly surprised at the performance of the transformer.

We'll close this discussion with a sample design. Let's say we want to transform approximately 5.5 ohms to 50 ohms over a frequency range of 2 to 50 MHz; the maximum power level will be 0 dBm.

The impedance transformation ratio is approximately 9:1, so a trifilar winding is required: $\sqrt{9} = 3$. With the relatively low power requirement, a small core is acceptable. I chose a Fair-Rite Products core no. 2843002802 on the basis that it had the smallest form factor. The number of turns to wind on the core is easily determined: To keep the transformer losses low, the minimum impedance presented to the lowest-impedance tap, with the secondary open circuited, should be approximately five to ten times that presented to the same terminals with the secondary loaded with its rated impedance. Since impedance presented to the lowest-impedance tap will be about 5.5 ohms when the total winding is loaded with 50 ohms, the impedance presented to the lowest-impedance tap with the load open circuited should be about 25 to 55 ohms. This calculation should be made at the lowest rated frequency of the transformer. When made for the lowest

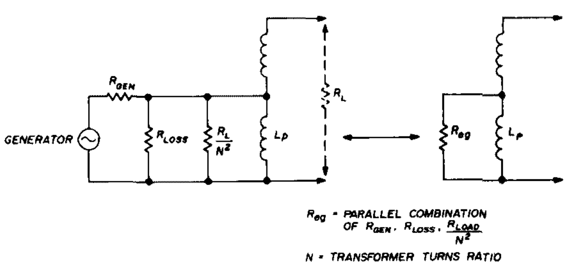
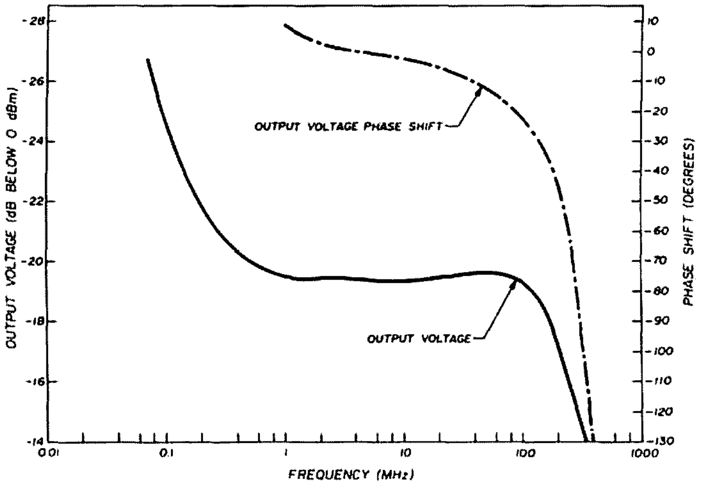


fig. 15. Derivation of the autotranformer primary inductance in terms of the total shunt resistance, which is the parallel combination of a) the load impedance referred to the primary, b) the parallel equivalent of the generator resistance, and c) transformer losses expressed as shunt resistance across the primary.

fig. 17. Response of the transformer shown in figs. 4 and 16. Load was a 5.6-ohm resistor across the lowest-impedance tap. Transformer input was from a 50-ohm generator. Output voltage in dB below 0 dBm and secondary winding phase shift are shown.



impedance tap, this calculation will determine the number of trifilar turns.

From the core manufacturers' literature for the core used, one turn will give a parallel impedance of about 30 ohms at 2 MHz, so that the transformer will require one (trifilar) turn. The three strands will be connected in series aiding. The circuit is shown in fig. 4 and a photo of the completed transformer is given in fig. 16. Fig. 17 shows the voltage across a 5.6-ohm resistor connected across the lowest-impedance tap when the transformer is driven from a 50-ohm generator. The measurements were made with a Hewlett-Packard 8405A Vector Voltmeter with a reference channel connected to the transformer input and the slaved channel connected across the load impedance.

The generator output was set to 0 dBm into a 10-dB, 50-ohm resistive attenuator. The left-hand ordinate in fig. 17 gives the output voltage in dB below 0 dBm. As the transformer has a 3:1 voltage step-down ratio, the output voltage would be expected to be -19.54 dB (-9.54 - 10 dB) below the input to the 10-dB pad (disregarding the impedance change). The right-hand ordinate gives the secondary voltage phase shift in degrees.

From fig. 17 it can be seen that the output is constant within a few tenths of a dB from 1 to 100 MHz. While this bandwidth is not as great as that obtained with transmission-line transformers, an impedance transformation ratio of 9:1 is obtained with one core and other impedance ratios are conveniently available, making the transformer useful for many applications.

fig. 16. Photo of the transformer design shown in fig. 4. Small size is key to good high-frequency response.

summary

The n-filar broadband autotransformer using only a single core gives greater flexibility than is possible with transmission-line devices, thereby achieving greater economy. Furthermore, autotransformers can be used in either balanced-to-balanced or unbalanced-to-unbalanced applications. While the autotransformer as described above may not be a cure-*all* for a circuit *designer's* problem, it certainly is a versatile component in his bag of tricks.

Those interested in experimenting with wideband transformers or baluns may obtain an assortment of the various cores shown in table 1 from Fair-Rite Products Corp.*

*Fair-Rite Products Corporation, Walkill, New York 12589. Price: $10.00 postpaid.

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ham radio

audio filters

for improving SSB and CW reception

The FX-60 active filter is put to work as a tunable lowpass filter for ssb and as a narrow bandpass filter for CW applications

Except in more expensive receiver designs, manufacturers have given little attention to the problem of processing detected audio. The usual practice is to roll off the higher-frequency audio signals using shunt capacitance or perhaps include an inductor to introduce reactance at the higher audio frequencies.

This article provides design and construction details for a lowpass filter for single sideband and a narrow bandpass filter for CW reception. The filters are designed around the FX-60 integrated circuit,* which is a cull from the manufacturer's product line but which is perfectly acceptable for the applications described.

*Kinetic Technology, Inc., 3393 De La Cruz Blvd., Santa Clara, California 95050.

improving receiver performance

To detect a single-sideband audio signal from the i-f carrier, some form of mixer is used with a bfo signal. The products resulting from this mixing consist of the audio signals plus an additional number of signals including the bfo frequency, the i-f, and a variety of complex sum and difference frequencies. Many of these signals are within the audible range and are heard as hiss.

The chief advantage of the single-sideband signal is its narrow audio bandwidth. By limiting the audio bandwidth within the transmitter and transmitting only one of the sidebands, we can concentrate the power where it's most effective, resulting in communicable information. The transmitted audio bandwidths will vary from 1.5 kHz to almost 3 kHz; in the process of limiting the transmitted signal bandwidth, normally both the very low frequencies and those above approximately 3 kHz are eliminated from the transmitted signal. The low frequencies are limited by using only nominal coupling capacitors between various audio processing stages and restricting the bandpass with a ceramic-, crystal-, or LC-filter section. Obviously, when the received signal is detected, the high- and low-frequency components contain no intelligence and can be eliminated. While restricting the received audio bandwidth to the useful range of information frequencies, we can improve the receiver overall performance and signal-to-noise ratio.

single-sideband filter

Fig. 1 shows a very simple lowpass filter for use with single-sideband receiver systems. A number of circuits are available in the various amateur handbooks for per-

By M.A. Chapman, K6SDX, 935 Elmview Drive, Encinitas, California 92024

forming this type of audio processing. Normally these circuits employ a series of toroidal-core LC sections in a variety of T or pi arrangements. This is a very effective scheme for single sideband audio filtering; however, to provide the necessary large inductive reactance, the

frequency tolerance over slightly wider ranges than the prime product-line devices, but in all other respects its performance characteristics are identical.

To minimize loading previous receiver circuit stages, the single-sideband filter circuit of fig. 1 employs a

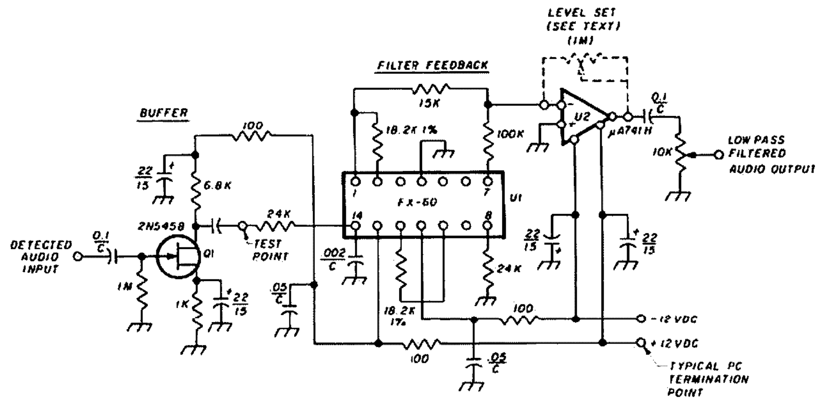


fig. 1. Schematic of the 2.5-kHz detected-audio lowpass filter for single-sideband application.

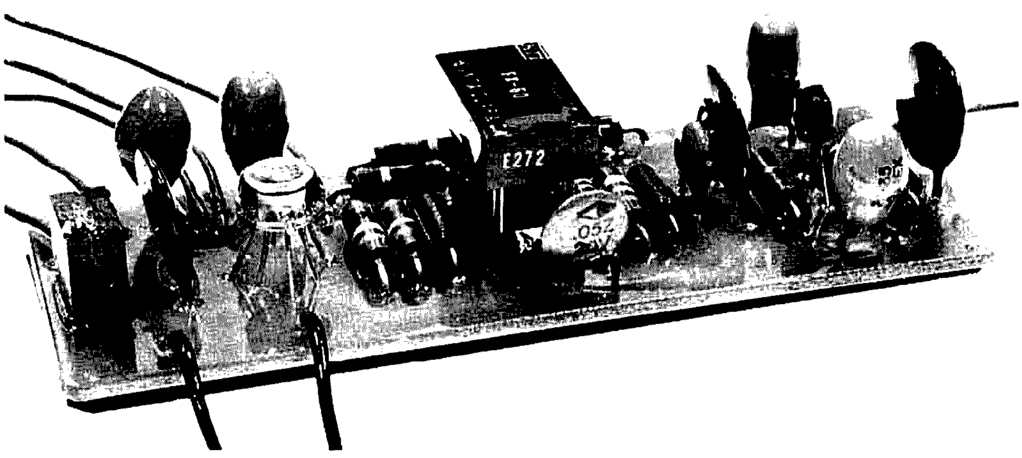
toroids and capacitors are physically quite large, and the performance curves indicate rather poor uniformity.

The circuit illustrated in fig. 1 uses an inexpensive hybrid active filter, which employs multiloop negative feedback for lowpass transfer functions. Tuning the filter frequency and selecting Q are easily accomplished by adding external resistors. U1, the FX-60 device used for the lowpass audio filtering, is a cull unit available in small quantities from the manufacturer, whose product line includes this device with very rigid parameters (and at a much higher price) under alternative part numbers FS-60, FS-65 and FS-61. The FX-60 will vary in center-

high-impedance buffer stage, Q1. This stage provides only nominal gain and isolates the filter from previous receiver stages. The lowpass cutoff or rolloff frequency is determined by the two precision resistors at terminals 2 and 10 of U1 (18.2k, 1%). Using the manufacturer's relationship for determining the lowpass rolloff point, we find that

$$f_c = \frac{455(10^5)}{18.2(10^3)} = 2.5 \text{ kHz (nominal)} \qquad \textbf{(1)}$$

where f_c is the filter cutoff frequency. This relationship results in a sharp rolloff of audio frequencies above 2.5



Construction of the CW bandpass filter which uses three active filter ICs. The bandwidth of this filter is 100 Hz, with a center frequency of 1000 Hz. All other audio frequencies are at least 20 dB down.

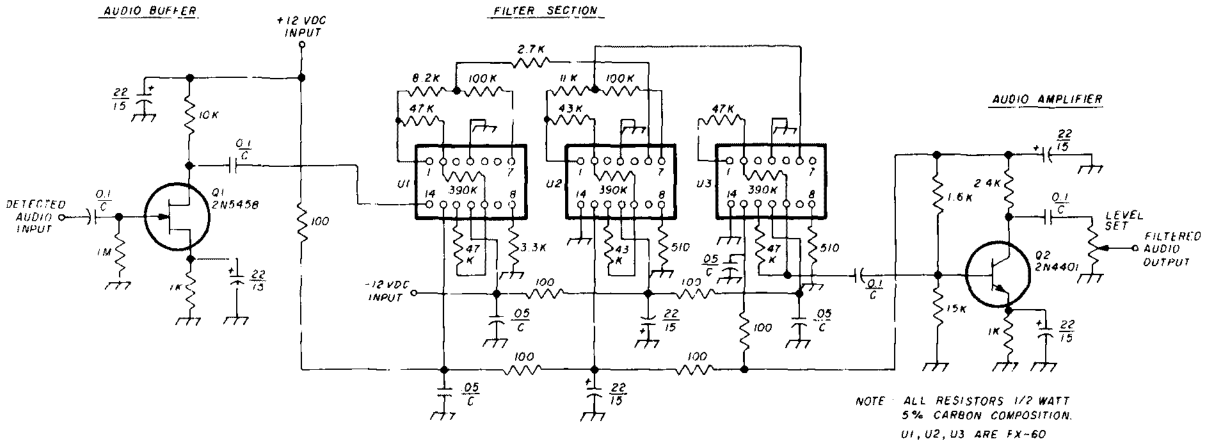


fig. 2. Schematic of the 900-Hz audio bandpass filter for CW. Bandwidth is ⩽ 150 Hz with minimum relative attenuation of ⩾ 20 dB.

kHz. A relative minimum attenuation of 15 dB is obtained within less than one octave; and as the frequency increases, this relative attenuation continues to increase. The overall gain of the single-sideband filter is achieved by including a simple operational amplifier, U2. The selection of a μA741 device here is based on its internal frequency compensation and stability over the bandwidth of interest.

CW filter

Fig. 2 is a somewhat more sophisticated audio-processing system for CW bandpass use in a receiving system. In CW reception we are normally interested in the audible quanties of the beat note resulting from mixing a bfo signal with an i-f carrier. Normally this beat note is clearest and most distinguishable at frequencies between 800-1000 Hz. Most manufacturers provide a bfo signal injection at approximately 1 kHz above or below the i-f for CW detection. This bfo signal is most often generated by a separate crystal or LC oscillator to maintain frequency stability.

Ideally we would like to pass the detected CW signal through some type of bandpass circuit that would allow only the 1-kHz beat note to be heard and that would discriminate against all other audio frequencies. Techniques for enhancing the CW beat note audio tones are described in several amateur handbooks and are similar to the single-sideband filter in that an LC system is used. These schemes rely on peaking the 1-kHz audio tone with tuned circuits of modest Q. Obviously, to provide tuned circuits at these low-frequency levels the L and C components must be physically large, which creates packing problems. Sharp bandpass filter characteristics are difficult to achieve using this LC filter method. To provide narrow bandpass features at 1-kHz would require that the LC system Q be quite high. Recall that the Q of a tuned circuit is

$$Q = \frac{X_L}{R} \qquad (2)$$

where $X_L = 2\pi L$ and R = reactance of the coil producing the inductance.

Since Q is proportional to L/R and both values are large for inductors usable at the CW beat frequencies, the LC filter is obviously difficult to implement and requires series-tuned circuits to narrow the audio passbnad.

The circuit illustrated in **fig. 2** uses the FX-60 in a multimode bandpass arrangement whose center frequency is ideally 1 kHz with the 100-Hz bandpass. It has a minimum relative attenuation greater than 20 dB for all other detected audio frequencies.

This ideal center frequency and bandpass width can't be achieved without considerable expense. First you'd have to use the FX-60 manufacturer's prime-line device, the FS-61; secondly, all the tuning resistors from pins 2 and 10 of U1, U2, U3 in **fig. 2** would have to be 1% precision resistors. The cost of such a system would be several hundred dollars. However, if you're willing to accept a compromise in the center frequency and a slight increase in bandpass width, a very high-performance CW bandpass filter can be built for one-tenth the cost of the ideal system.

Normally the FX-60 output will fall on the low side of the manufacuter's nominal center frequency, which is the reason it's a cull unit. Using the resistance values indicated in fig. 2, the center frequency would be ideally 1 kHz. Because the FX-60 units are on the low side, the actual bandpass center frequency will be between 900-950 Hz. Also because of the variations between units and the tolerance of the 5% carbon composition resistors used for tuning on pins 2 and 10 of U1, U2 and U3, the bandwidth will increase from the ideal (100-Hz) to approximately 150 Hz. As with the single-sideband filter, a high-impedance input buffer (Q1) is used, operating at unity gain. Bandpass signal amplification is achieved using a simple class-A common-emitter amplifier, Q2.

An improvement in bandwidth can be achieved in the CW filter by using matched pairs of resistors from pins 2 and 10 of U1, U2 and U3. How much improvement, of

course, depends on the accuracy of your measurements and on your patience in selecting these resistors.

construction

The photos illustrate the construction of both the single-sideband and CW filters. Each is assembled on a single-sided PC board, with all components mounted on the side opposite the etched foil pattern. **Fig. 3** illustrates the relative component placement and all input-output points. Generous use is made of isolation resistors and bypass capacitors for both the plus and minus 12 volt bus lines of the ssb filter unit.

If the lowest practical information bandwidth is approximately 250 Hz, then the 0.1-μF interstage coupling capacitors are more than adequate, considering the high input impedance of successive stages. This same reasoning applies to the power bus decoupling capacitance (22 μF). You can substitute lower-value capacitors freely without serious performance effects assuming, of course, that the power bus ripple content is reasonably low. The feedback control resistor for U2 is shown as a potentiometer; however, after assembly and test, a fixed resistor may be mounted permanently to the board in the space provided.

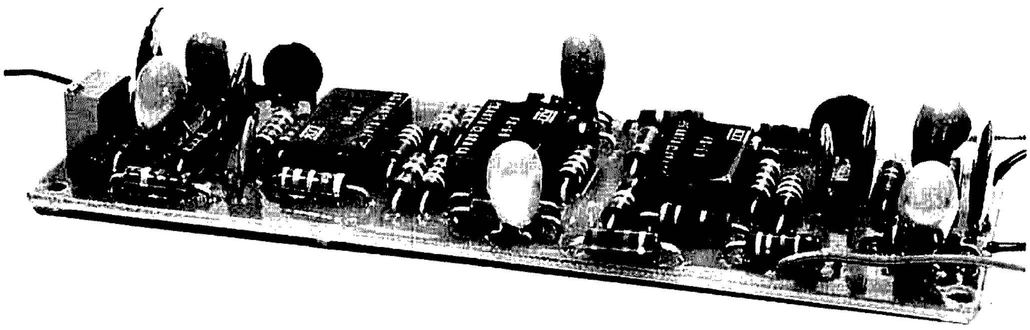
Fig. 4 shows component placement for the PC assembly of the CW filter. Note that the FX-60 is mounted to an IC socket for the single-sideband filter and directly to the board for the CW bandpass unit. I did this purposely to illustrate that either assembly method is acceptable. IC sockets are used to allow easy device removal should they be faulty at the time of initial installation or fail during use. These units are culls, so you can be assured they've been thoroughly tested, thus there's little chance they won't function initially. Because of their conservative design, these devices have an almost indefinite life and a very small possibility of infant mortality.

Resistors were chosen to reduce noise generation and not for power dissipation. Total power dissipation for both filters is in the order of a 100 mW, being concentrated at the output stages of both assemblies. You may substitute ¼ watt or smaller resistors without concern of thermal restrictions.

hints and kinks

Many n-channel depletion-mode, audio-type fets may be used as substitutes for the devices in the Q1 buffer stage. The 2N5458 has a modest I_{DSS} value and if alternative devices are used, they should have the same general ratings. Units such as the 2N5459, which has a higher I_{DSS} value, will require that the source resistor be reduced to approximately 500 ohms and that a lower-value drain load resistor be used. This drain resistance value can be obtained by temporarily connecting a pot in the drain circuit and adjusting the quiescent voltage for approximately 6 volts at Q1 drain termination. Remove the pot and install a resistor with a value closest to that measured on the pot. The CW filter output stage is a simple class-A common-emitter amplifier. Any similar circuit arrangement and device types with which you're familiar can be used, since the circuit is capacitively coupled.

Two important considerations must be kept in mind in the design of the Q2 stage. First, the FX-60 output impedance is in the order of 100 ohms and to develop maximum output voltage to the base of Q2, its input impedance should be 500-1000 ohms minimum.



Construction of the ssb lowpass ssb filter which uses a single KTI FX-60 active filter IC. Operational amplifier (in TO-5 package, left) provides overall gain.

*Undrilled boards are available from the author for $3.00 each postpaid.

Secondly, in the output stage, the ac load consists of the collector resistor and the output coupling capacitor reactance together with the level-set potentiometer in parallel with the signal. If your favorite class-A circuit is used here, some adjustment may be required in the collector resistance and base bias to achieve maximum signal output levels as a result of the added loading.

Never apply power to the FX-60 without the external resistors attached to pins 2 and 10, or the internal circuits will latch and functional properties will be destroyed.

The output-level adjustment allows the filters to be used either as unity gain or preamplification gain stages. Where preamplification is not required, the feedback-level set on the single-sideband lowpass filter should be adjusted for about 80% or less of the maximum voltage gain available. Adjustment at levels slightly below maximum will ensure good audio-signal stability, low noise, and minimum distortion.

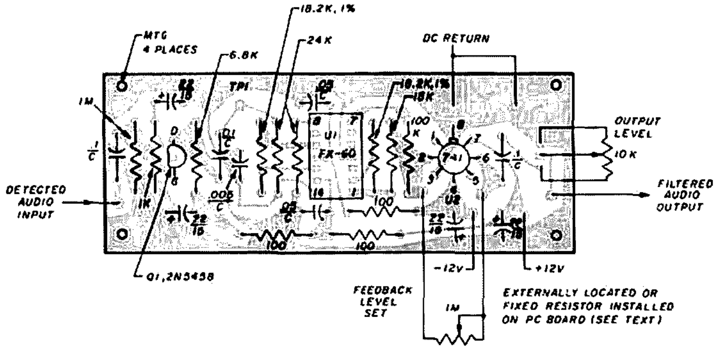


fig. 3. Printed-circuit layout for the single-sideband lowpass filter.

application

Table 1 summarizes the performance of both circuits. Both filters provide considerable voltage gain, which may be used to supplement or replace audio preamplification stages in your receiver. Usual receiver detection systems have audio output levels between 10-100 millivolts, depending on the type of circuits used, with subsequent audio amplification. As shown in table 1, the maximum peak-to-peak output-voltage swing is greater than 6 volts, indicating that the filter may be used at any point within the receiver audio system between the detector and final audio amplifier.

When the CW filter is used in receivers having crystal bfo oscillators, it is necessary to first assemble the CW filter and test its performance using an external variable audio oscillator to determine the center frequency. This frequency will usually fall between 900-950 Hz.

After determining the bandpass-filter center frequency, a bfo crystal must be used whose frequency corresponds to the receiver i-f and the filter center frequency. As an example, suppose your receiver i-f is 455 kHz and the measured center frequency of your CW bandpass filter is 920 Hz. The correct bfo crystal frequency is then X_f = 455 kHz±920 Hz. You have two choices: 455.92 kHz or 454.08 kHz. Therefore the i-f carrier can be heterodyned from 920 Hz above or below the nominal center i-f to obtain the desired CW audio signal.

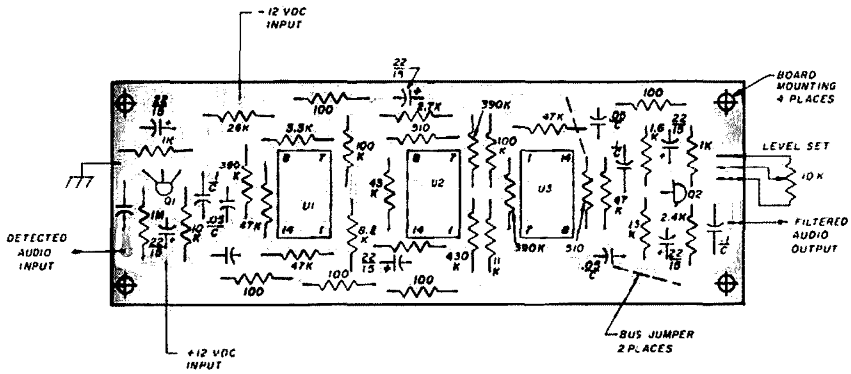
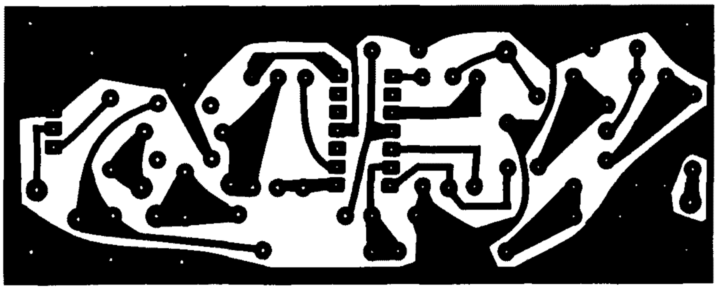


fig. 4. CW filter printed circuit board component assembly.

It is not uncommon for many receiver manufacturers to use a single-sideband crystal bfo frequency for CW reception. This is entirely reasonable, since the i-f pass band might be in the order of 2.5 kHz wide. The upper- or lower-sideband crystal is chosen to beat with the single-sideband i-f signal at ±20 dB from the i-f center frequency. These bfo frequencies will usually be slightly above 1 kHz; perhaps 1.3 kHz in the nominal case, which is adequate for CW detection. However, these frequencies result in audio tones slightly higher than desired. It is not possible to use the CW filter circuit described here with this type of detection schème. You must either alter your receiver bfo crystal frequency or use resistance values in the tuning of U1, U2 or U3 to move the center frequency up to correspond with the receiver crystal beat frequency. This can be done using the curves furnished by the FX-60 manufacturer and trimming the resistance values to offset the center-frequency tolerance of the particular FX-60 device being used.

The filters can be wired so that each is selectable by a panel switch, or each filter can be installed as an integral part of your receiver without a switch. For CW, the filters can be cascaded to attenuate high-frequency audio noise. In any event, the filters should be wired to avoid ground loops and oscillation. Isolate the PC-board ground plane from the receiver chassis and allow the dc return to enter the board at only one point.



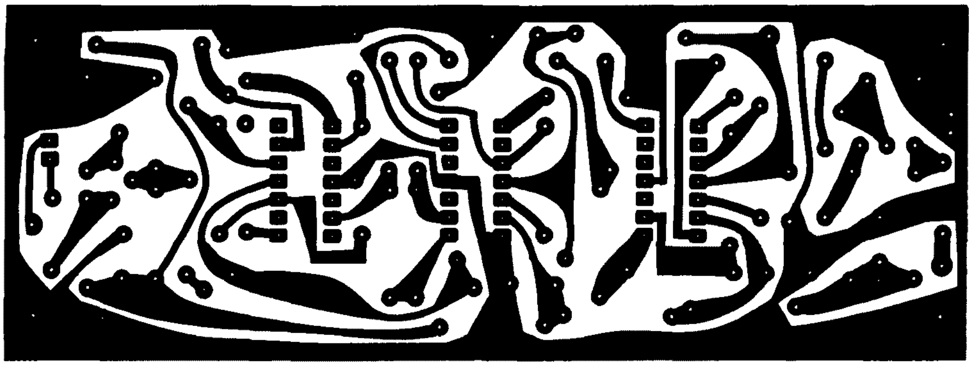


fig. 5. Full-size printed-circuit boards for the single-sideband lowpass filter, above, and the CW filter, below.

Long leads carrying the input signal to the filter should be shielded. Ground the shield to the filter PC-board ground plane near the Q1 input and leave the other end of the shield open. The output signal may or may not be shielded, depending on the filter location, its lead length, and signal level. If output shielding is used, the shield should be terminated at the input of the subsequent stage, and the filter output shield should be unterminated. These wiring methods will avoid the possibility of introducing a ground loop and will reduce external noise pickup.

Applications of both filter units with receivers using LC bfos may be optimized by adjusting bfo frequencies to correspond to a) the filter pass band cutoff frequency in the case of the lowpass unit and b) the filter pass band center frequency for CW filter. In most cases this can be accomplished by adjusting the single-sideband bfo frequencies for maximum clarity of the received signal and by adjusting the CW bfo for maximum amplitude of the best tone.

ham radio

table 1. Performance summary of the receiver audio filters for CW and single-sideband.

parameter	CW filter	lowpass filter
maximum voltage gain (note 1)	⩾30	⩾100 (50 nominal)
center frequency	1 kHz ±10% nominal	not applicable
rolloff frequency	N/A	2.5 kHz ±10% nominal
1st octave relative attenuation	>20 dB	>15 dB
noise (note 2)	<15 mV	<10 mV
maximum peak-to-peak output signal	⩽ 9V	⩽ 10 V

(1) Voltage gain of CW filter may be increased by device and value selection of Q1 stage. Maximum possible Q1 voltage gain ≈ 10; nominal voltage gain of Q2 stage ≈ 10.
(2) Actual noise generation depends on source-voltage ripple content indicated relative to maximum p-p output signal with input grounded.

200 meters and up

receiving converter for low frequencies

With this two-transistor converter you can explore the fascinating world below the a-m broadcast band

The story of early amateur radio has been told in *200 Meters and Down*, a book authored by Clinton B. DeSoto in 1936 and published by The American Radio Relay League [1] It makes interesting reading and establishes the background for the scientific hobby of amateur radio as we know it today.

More than fifty years have passed since radio amateurs abandoned the very-low frequencies and started to explore the spectrum above the a-m broadcast band. Just what are the longer wavelengths being used for today? The only way to find out is to listen. Considering that the World Administrative Radio Conference will be held in Geneva, Switzerland, in 1979 and that the WARC-79 task group in this country is considering a proposal for an amateur band in the 150-200 kHz region,[2] it seems appropriate that the experimentally inclined amateur should be exploring the low-frequency radio spectrum.

This article presents construction information for a simple solid-state converter that can be used with any good-quality communications receiver to tune the frequency spectrum below the a-m broadcast band.

frequencies available for communications

The FCC permits the use of *nonlicensed* transmitters in certain parts of the medium- and low-frequency radio spectrum. Section 15, paragraph 15:203, permits the use of a 1-watt transmitter and a 50-foot (15.2m) antenna* between 160 and 190 kHz. Paragraph 15:204 permits the use of a 100-mW transmitter and a 10-foot (3m) antenna* in the range 510-1600 kHz, which includes the a-m broadcast band in the United States. Any transmission mode can be used on these frequencies as long as any emissions outside the band edges are suppressed 20 dB below the unmodulated carrier.

background

For the past six years I've been experimenting with these frequencies for communications use, and as of this writing my 189.7-kHz beacon "K" is regularly copied by an amateur in Riverhead, Long Island, New York, some 90 miles distant. My 1575-kHz beacon is copied by

*Antenna length also includes the length of any feed line.

By Ken Cornell, W2IMB, 225 Baltimore Avenue, Point Pleasant Beach, New Jersey 08742

another amateur in Lincroft, New Jersey, 18 miles away. Not bad for QRP_p with mini-antennas!

For listening on the medium-, low- and extra-low frequencies, a good receiver is desirable. While there are many usable surplus receivers on the market, they still command a fair price and in many cases require conversion surgery. Converters for vhf and uhf are common, and the same basic heterodyne logic can be applied for the lower frequencies (see **fig. 1**).

The main purpose of this article is to review and update a converter described in reference 3. This converter uses two pnp transistors, one as a mixer and the other as a crystal-controlled local oscillator. It's designed to work with a receiver tuning the 80-meter band and uses a 3500-kHz crystal.

In this case, 10 kHz will appear at 3510 kHz and 500 kHz at 4000 kHz on the receiver dial. A 1-kHz frequency readout or better, depending on the receiver tuning accuracy, can be obtained through the tuning range. The only coil that must be changed is the antenna-to-mixer input coil. This coil and its tuning capacitor must provide a resonant circuit through the converter tuning range.

My Riverhead, Long Island, New York anchor man uses this converter with his Drake R4-A receiver, and his reception reports of experimenters using the 160-190-kHz band are outstanding. My receiver for the lower frequencies is an HRO5TA1. I have all the coils for reception from 50 kHz to 30 kHz. I also use a Central Electronics model DQ Q-Multiplier and a Heath panadaptor as accessories.

My type of experimentation required a receiver that I could use for portable work with a 12-volt battery, so I duplicated this converter and made minor revisions so that it would work with my Yaesu FT-101 transceiver. I was quite pleased with its performance. I am also making some experiments in the 7-10 kHz range, which the HRO doesn't cover, and the converter does a good job there as well.

tuning capacitor

The only complicated construction for this converter, as detailed in reference 3, is the rf tuning capacitor. It calls for two 3-gang variables with the stators wired in parallel. The two capacitors are mounted side-by-side using dial cords for tuning. Such construction could tax the patience of the most experienced amateur.

This problem can be resolved in two ways. One is to obtain a 5-gang surplus variable capacitor.* These capacitors appear to have a total capacitance, with the stators wired in parallel, of 2000 pF. The capacitors I obtained have a 3/8-inch (9.5mm) diameter shaft. I purchased some 3/8-inch (9.5mm) female-to-¼-inch (6.5mm) male shaft reducers from Lafayette Radio (part no. 32-64165) to accommodate a common dial to mount the capacitor behind the panel.

*I obtained mine from Ralph Sanserino, 8422 Crane Circle, Huntington Beach, California 92646. A similar capacitor is offered by Fair Radio Sales, P.O. Box 1105, Lima, Ohio 45802.

The other way is to use the largest variable capacitor you can obtain. If it's the common BC-set type, a rotary switch can be used to shunt fixed capacitors across the variable capacitor to increase its range, otherwise many more coils will be required through the 10 to 500 kHz range. An estimate is that, with a 400-pF variable, you'd need about a dozen coils in this case.

improved converter

I used the converter as is for some time but decided to make some circuit changes to permit more flexibility.

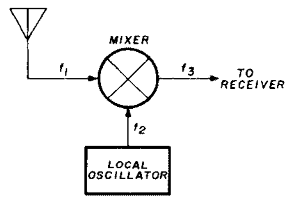


fig. 1. Basic converter heterodyne logic. f_1 **is the desired received-signal frequency,** f_2 **the local-oscillator frequency.** $f_3 = f_1 \pm f_2$, or $f_1 \pm f_3 = f_2$.

To achieve this flexibility, I changed the local-oscillator circuit to a Pierce type and used a Motorola HEP-802 fet transistor, which eliminated the need to change the oscillator coil to suit the crystal frequency. The mixer output coil is a plug-in type. The circuit of my revised converter is shown in **fig. 2.**

construction

Because of the large size of the rf tuning capacitor, which is 6 inches long by 3½ inches high by 2½ inches wide (15x9x6.5cm), all thoughts of miniaturization were abandoned. I made a plywood box 10½ inches wide by 8 inches deep by 8½ inches high (27x20x22cm) that contains the converter, which is built on a baseboard. The panel is black plastic, and the cover is hinged for access to the interior and also has a carrying handle. Except for the input tuned circuit, all components are mounted on a small piece of copper-foil board located on the right-hand side of the rf-tuning capacitor, which is centered on the base. Sufficient space is available on the baseboard and panel for mounting coils and switches. The space on the left-hand side of the rf tuning capacitor contains an audio filter for extreme selectivity.

The secret of success with this converter is the antenna-to-mixer input tuned circuit (C2, L1, C3), which is a pi network. It must be resonant at the received-signal frequency for maximum sensitivity. Thus constant peaking of capacitor C2 is necessary as you tune through each L1 coil range.

coils and switching arrangement

The simplest method for mounting L1 is to make it a plug-in coil, or a more complex method is to mount the coil on a suitable base and wire each pie to a rotary switch on the panel. I used a combination of these methods in the converter shown in **fig. 2.**

On a coil-winding machine I wound a series of five pies on a 3/8-inch (9.5mm) diameter dowel.* Each pie

contains 50, 100, 150, 200, and 300 turns of 18/42 Litz wire. The winding is continuous and a tap is provided between each pie.

I used a two-gang, eight-position rotary switch wired in the following manner for L1: One gang of the switch was used in the first five positions for this coil. The sixth position was wired to a plug-in coil socket, and the seventh position was wired up through the second gang of the switch to a pair of binding posts. These binding posts permitted the addition of fixed capacitors across the plug in coil, assuring the ultimate in tuning flexibility, especially in the extra-low-frequency ranges.

A suggestion for plug-in coils, if desired, is to mount them on a dual banana plug, a common item available from most radio-part suppliers. These plugs can be stacked, so that you can use one plug with a capacitor connected across the pins, and when plugged into the permanent socket, the plug containing the coil can be plugged into the capacitor plug. This makes for a flexible arrangement to obtain various LC ratios.

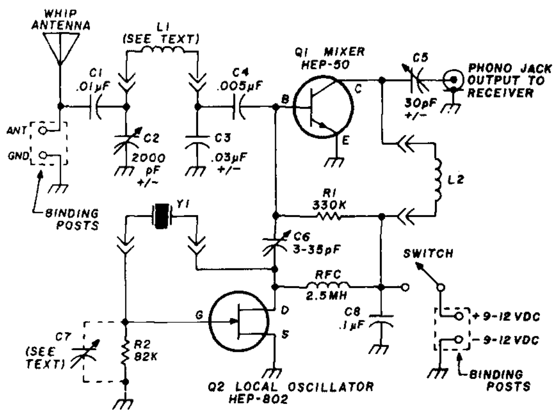
checkout

After the converter has been assembled and checked for wiring accuracy, use your vom to check coil continuity and possible short circuits, such as a bad bypass capacitor. Apply power and check the local-oscillator frequency on your receiver for reliable operation. In my case, a sluggish 3500-kHz crystal (FT-243 type) would start oscillating only when I pinched the crystal between my fingers – I later found that if I placed a small value of capacitance between the Q2 gate and ground, the crystal would trigger right off. I used an Arco/Elmenco no. 120 (1-12 pF) and adjusted it for optimum performance (This *capacitor* is C7 *in* **fig. 2**).

The number of stations that can be heard on "200 meters and up" are too numerous to list here, but the low end of the a-m broadcast band will provide signal sources to check out the medium frequency portion of the converter. For low frequency, look for WGU-20 on 179 kHz. This is the first of several planned civil defense preparedness stations. It is located in Chase, Maryland, and gives Eastern mid-Atlantic weather reports and accurate time. There is also TUK on Nantucket Island (194 kHz) and SFI in San Francisco on 192 kHz. On extra-low frequency, various Omega navigation stations are scattered around the world (13.6 kHz).

in conclusion

The major problem when using a converter with an auxiliary receiver is fundamental frequency feedthrough. This is because the receiver is trying to do the job it was designed for, and if the converter is not well shielded and an unshielded wire connection between converter and receiver is used, feedthrough can occur if strong stations are in your area. While I have had no serious



C1	0.01 µF (most any type)
C2	2000 pF variable (see text)
C3	0.02 to 0.04 µF disc ceramic (experiment with value for best results)
C4	0.005 µF disc ceramic
C5	Arco/Elmenco 422 trimmer or equal should be adjusted for best performance
C6	Arco/Elmenco 403 trimmer or equal should be adjusted for best performance
C7	see text
C8	0.1 µF disc ceramic
L1	target value for inductances (See text): 5-11 kHz 0.28 H 50-100 kHz 3.5 mH 10-20 kHz 100 mH 90-200 kHz 1 mH 18-38 kHz 25 mH 150-350 kHz 350 µH 30-70 kHz 10 mH 250-550 kHz 120 µH
L2	80-90 µH for 80 meters. Loop stick for BC band
Q1, Q2	Motorola transistors, but substitutes will work as long as npn tr. is used for Q1 and an fet for Q2
R2	50-100k; 82k nominal
Y1	3500 kHz is used for the 80-meter band but other crystals can be used to suit your heterodyne logic (fig. 1)

fig. 2. Schematic of a medium-, low-, and extra-low-frequency converter for use with an amateur communications receiver.

problem on 80 meters, I believe I'll have to line the inside of my wood cabinet and plastic panel with aluminum foil if I use the converter with a broadcast receiver.

A wave trap can also be used to attenuate an interfering signal. This trap is nothing more than a coil with a variable capacitor across it. The trap is placed in series with the antenna and is located as close as possible to the converter.

references

1. Clinton B. DeSoto, *Two-Hundred Meters and Down,* American Radio Relay League, 1936 (reprints are available from the ARRL).
2. "Presstop," *ham radio,* November, 1975, page 6.
3. William H. Fishback, W1IKU, "A VLF Converter for Communications Receivers," *QST*, September, 1968, page 18.

ham radio

*The Morris hand-operated coil-winding machine, complete with all gears and cams, is available from Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, Long Island, New York 11791. Catalog number 32-F-87018.

electronic bias switch for negatively biased amplifiers

This circuit is designed for the Heath SB-200 but can be adapted to any linear amp using negative bias voltages

Several excellent articles have appeared recently in the amateur literature describing automatic electronic bias switching of linear amplifiers. Bryant[1] published an article in *QST* describing the use of electronic bias switching in the ETO Alpha 77 amplifier. Also included was his adaptation of the circuit for use with the Heath SB-220 linear amplifier. Gonsior[2] published an article in *ham radio* describing his refinements of the Bryant circuit to allow for controlled rise and fall times to create a softer switching action on the bias circuit.

Realizing the importance of electronic bias switching for conservative amplifier operation, I adapted these techniques to the Heath SB-200 linear amplifier, which uses negative rather than positive bias voltages. Amplifier efficiency is enhanced using electronic bias switching, because no power is dissipated under no-signal conditions.

system operation

The Heath SB-200 is designed for linear class-B operation. During the transmit mode the amplifier output tubes, type 572B, are biased with −2 volts on the grids. This bias voltage allows 90 mA of plate current to flow under no rf drive conditions. With 2400 volts on the 572B plates, the quiescent power consumption is nearly 240 watts continuous. Thus the tubes must dissipate this power under no-drive conditions, creating heat that doesn't contribute to amplifying action. If the tubes are completely cut off when no rf drive is present the plate current would be zero. Hence, the power dissipated would be zero under no-signal conditions. This is the purpose of the electronic bias switch.

Fig. 1 is a block diagram of the switch, along with the SB-200 bias circuit. The SB-200 uses negative grid voltage to bias the tubes, whereas the SB-220 uses positive cathode voltage to bias the amplifier for linear operation. The electronic switch senses the presence of rf drive voltage and switches on the class-B bias voltage only when drive is present. With no rf drive, the tubes are cut off by a large negative voltage, and plate current ceases to flow. The electronic switch is very fast and responds to very small rf input voltages. By introducing a small delay into the switch action, a softer on-off action can be created, which results in a softer sound at the receiving end.

Fig. 2 shows the electronic switch circuit used with the SB-200. The circuit is very similar to that described by Gonsior.[2] There are two major differences, however, between this circuit and those published previously. First, negative voltages are controlled in the SB-200 rather than positive bias voltages. Thus, the circuit must be connected in an opposite manner to the previous designs. Second, the circuit must be connected so that the rf input has a good rf path to ground. This is the reason for the 0.001 μF capacitor across the transistors. The negative voltages switched are about −150 Vdc maximum. The transistors were chosen to withstand

By F. E. Hinkle, WA5KPG, I/O Engineering, 9503 Quail Trail, Austin, Texas 78758

these voltages without the need for additional crowbar protection as developed by Bryant.[1] Almost any high-voltage transistors may be substituted if care is used to ensure that the maximum voltage ratings are not exceeded. These transistors are Motorola devices and cost less than $2.00 new. The capacitor across the collector-to-base junction of Q1 is an integrating capacitor in the Miller feedback principle. This capacitor value may be adjusted to reduce the turn-on time of the switch.

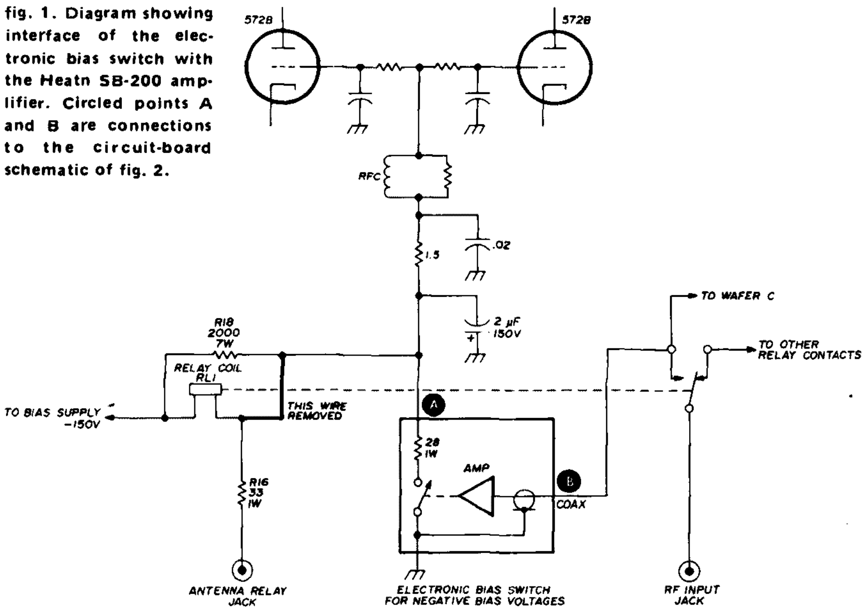


fig. 1. Diagram showing interface of the electronic bias switch with the Heatn SB-200 amplifier. Circled points A and B are connections to the circuit-board schematic of fig. 2.

SB-200 modifications

The photo shows the electronic bias switch as installed in an SB-200 amplifier. The location of the circuit board was chosen to permit the use of one of the wires removed during modifications. Using the following procedure, only two wires need be removed from SB-200 terminals and one wire added. No holes were drilled into the chassis, and the original circuit can be reconnected in a matter of minutes.

The circuit of fig. **2** was built on a small piece of Vector board. The board was then attached to a piece of aluminum strip about 2 inches (51mm) longer than the circuit board and formed as shown in the photo. The aluminum strip was then attached to one of the transformer-mounting screws protruding below the chassis. A new nut was used to fasten the aluminum strip to the bolt. The yellow wire from the rf relay coil was then removed and attached to the pair of **56**-ohm resistors as shown in fig. **2** at *point A*. The removed yellow wire was located on the relay coil terminal nearest the edge of the chassis (lower left corner of the photograph).

The yellow wire in series with the 33-ohm **1**-watt resistor located at the *Ant Relay* jack was removed next (see fig. 1). A piece of spaghetti tubing was placed over

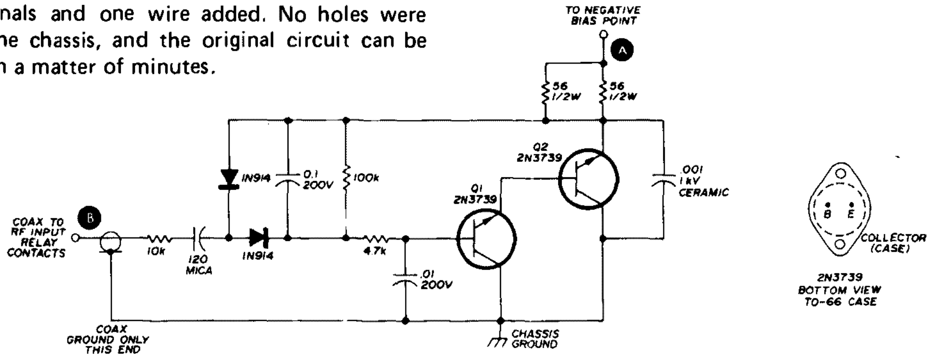
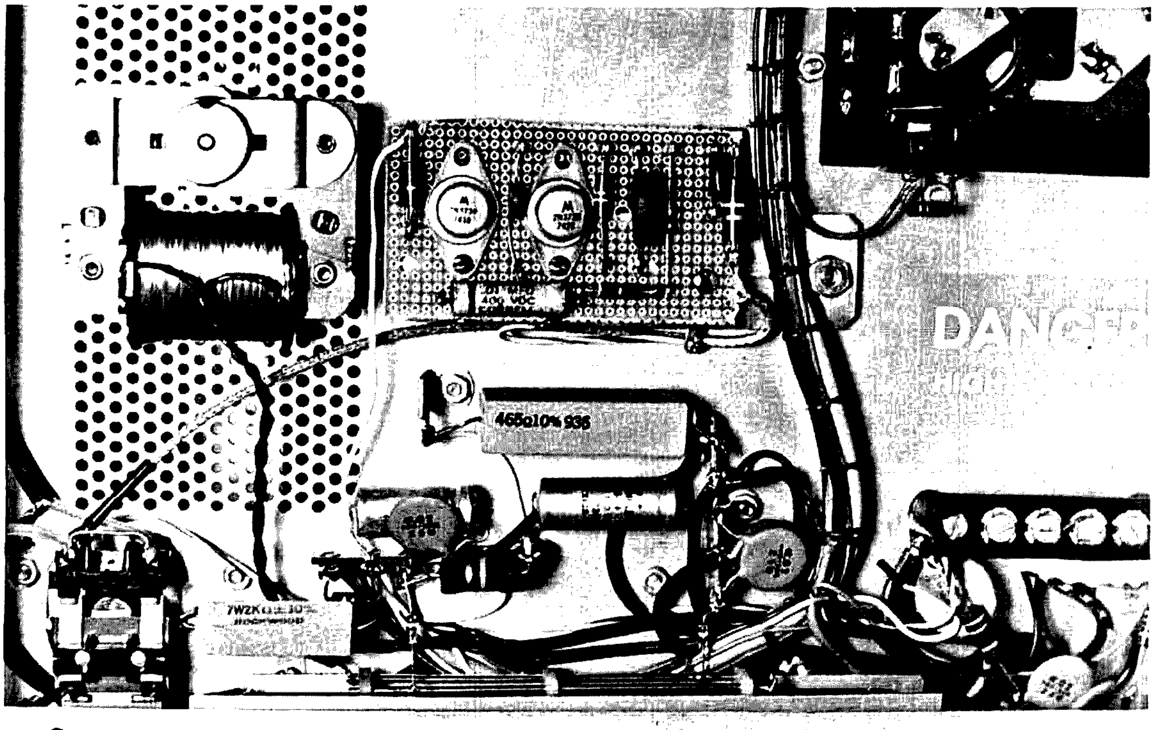


fig. 2. Schematic of the electronic bias switching circuit for the Heath SB-200 amplifier. Components are mounted on a piece of perf board, which fits into an unused space below chassis. No holes need be drilled. Only two wires are removed from terminals in the SB-200 and one wire added to make the modification.



Underchassis view of the SB-200 showing addition of the electronic bias switching circuit. Circuit board is mounted onto an aluminum strip, which is bolted to one of the transfromer mounting studs.

this yellow wire since it isn't needed. A new wire was then soldered onto the terminal from which the second yellow wire was removed. This new wire was then attached to the relay coil terminal just vacated. The *Antenna Relay* jack will now operate the rf relay coil but not the bias circuits.

A small piece of coax was then soldered between the circuit board and the rf relay contacts (*point B,* figs. 1 and 2). These contacts will have the *input* rf present on them when the relay is energized. The relay lug to which the coax was connected was the one nearest the top of the photograph, second from the top of the relay. Rf energy should be fed to the circuit board only if the relay is energized. The circuit board ground wire was attached to the aluminum strip but could be attached to any convenient ground lug.

operation

Proper operation of the electronic bias circuit is easily identified. Turn on the amplifier. If there is no rf drive from the transmitter, the amplifier plate current should be zero, indicating that the amplifier is biased to cutoff. Place the transmitter in the *tune* position. With the small amount of rf present at the output, the electronic bias switch will detect the rf and apply class-B bias voltage to the amplifier. The plate current should be 90 mA. As the rf drive is increased the plate current will increase as usual. When rf drive is removed the plate current should again decrease to zero. The amount of rf-drive voltage necessary to enable the bias switch is about 2 volts.

When the transmitter is placed in the ssb mode, with no speech, the plate current will be zero. With speech the plate current will increase in accordance with the rf driving voltage.

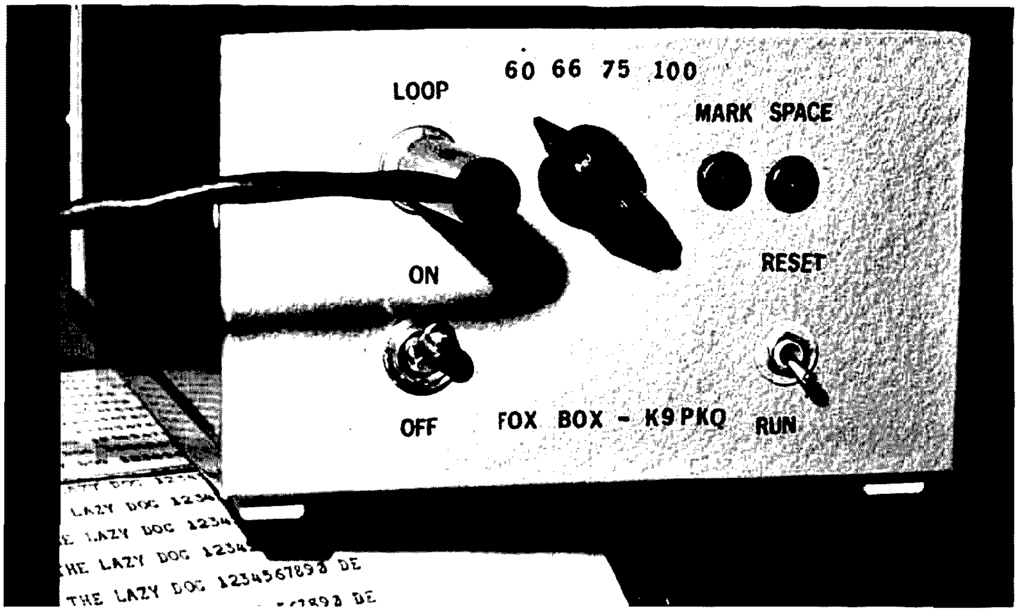
The electronic switch will respond to very small voltage levels, so it's mandatory that the ssb carrier be suppressed sufficiently. If an indication of amplifier plate current is present in ssb mode with no speech, the transmitter balanced modulator should be checked for proper carrier suppression in accordance with the manufacturer's specification.[3]

Although the circuit modification was for the SB-200 linear amplifier, there is no reason why it would not function in other amplifier designs which use negative voltages for biasing. References 1 and 2 should be consulted, however, for more insight into the operation of the bias switch and possible applications to other amplifier designs.

references

1. J. A. Bryant, W4UX, "Electronic Bias Switching for RF Power Amplifiers," *QST,* May, 1974, page 36.
2. M. Gonsior, W6FR, "Electronic Bias Switching for Linear Amplifiers," *ham radio,* March, 1975, page 50.
3. *Heathkit Assembly Manual, SB-200,* Heath Company, Benton Harbor, Michigan.

ham radio



RTTY test-message generator

Tired of fighting the cumbersome old fox box? Here's a compact, lightweight replacement using TTL logic and a mos read-only memory chip

RTTY buffs are familiar with the "fox box", a large, heavy electromechanical machine complete with commutator that generates the test message, "THE QUICK BROWN FOX JUMPS OVER THE LAZY DOG 1234567890." The disadvantages of this monster are well known, but when it came to testing a TTY machine for each function, the old fox box was a necessity.

This article describes a successor to the fox box that weighs only a couple pounds and is designed for the four popular RTTY speeds of 60, 67, 75 and 100 wpm. It generates "THE QUICK BROWN FOX JUMPS OVER THE LAZY DOG 1234567890 DE." Someone more ingenious could probably figure out a way to insert a call sign; this feature was not included, because this unit is normally used on a local loop with no requirement for a call sign.

A schematic of the test generator main frame is shown in **fig. 1**. It is designed around the MM5220DF "quick brown fox generator," which is one of a series of preprogrammed read-only memory (*rom*) ICs by National Semiconductor, Inc. All logic in the test generator is TTL except for the MM5220DF IC, which is a mos device. All parts can be easily obtained except the

By Ken Ebneter, K9GSC, and Jim Romelfanger, K9PKQ. Mr. Ebneter may be reached at 117½ 4th Street, Baraboo, Wisconsin 53913; Mr. Romelfanger's address is 822 Wauona Trail, Portage, Wisconsin 53901.

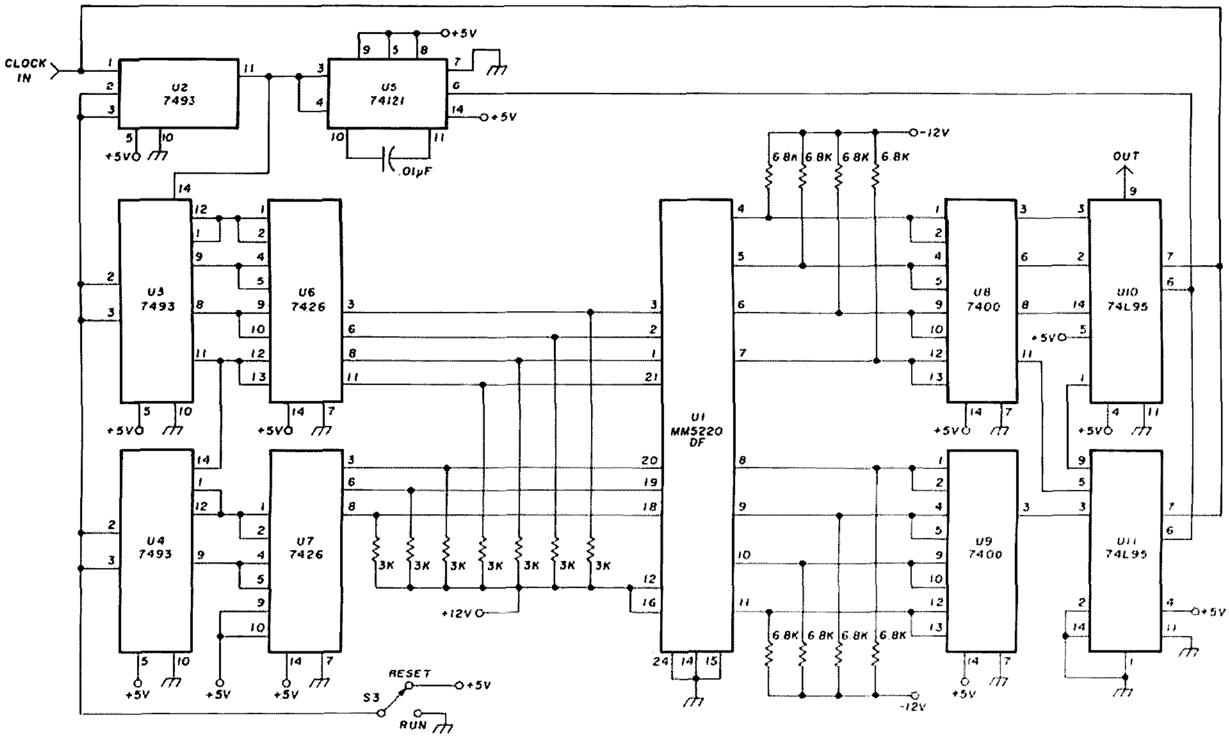


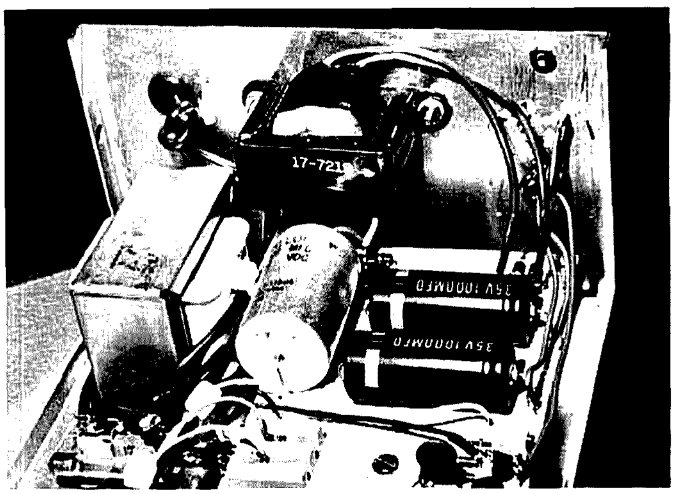
fig. 1. Schematic of the solid-state replacement for the old electromechanical "fox box." Design is based on National Semiconductor's "quick brown fox generator," a mos read-only memory chip. The other devices are TTL ICs.

MM5220DF. This device can probably be obtained from any National Semiconductor distributor or from Taylor Electric Company.*

circuit description

clock. This simple circuit was burgled from the Micro-TO Keyer[1] with some obvious modifications, fig. 2. It delivers a very sharp negative-going pulse train, which is TTL compatible. Four series resistors and 10K trimpots set the various baud rates. The clock can be adjusted for the proper speed by connecting a frequency counter to the clock output and setting the four trimpots for the desired baud rate as chosen by rotary switch S2; i.e., 45.45 for 60, 50.0 for 67, 56.88 for 75, and 74.2 for 100 wpm.

View showing power-supply components.



main frame. The clock drives the binary counter chain consisting of U2-U4. U2 is connected as an 8-bit divider. In addition to feeding the remainder of the divider chain, U2 output is also fed to U5, a one-shot that provides a load pulse for output shift register U10, U11. Counters U3, U4 provide the six address lines for the mos IC. The outputs from the counters must first be changed from TTL level (positive logic) to mos-compatible levels (negative logic). This is done by using gates U6, U7 and 3.0k pullup resistors. For each of the 64 different addresses, a different output word will occur, consisting of a Baudot letter or function. The mos output is buffered back to TTL level by gates U8, U9.

Parallel data from the memory is fed to shift registers U10, U11 along with a hard-wired start pulse (space) and

*Taylor Electric Company, Industrial Sales Division, Post Office Drawer 11N, Milwaukee, Wisconsin 53201. (Latest price quote is $18.00).

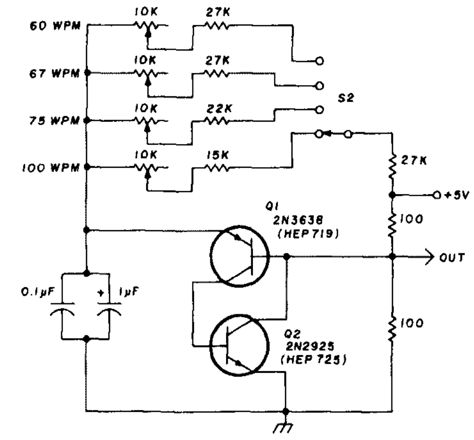


fig. 2. Clock/circuit schematic for the RTTY test-message generator. Desired baud rate is selected by the rotary switch for the four popular RTTY speeds.

two stop pulses (mark) to be run out in serial form by the clock pulses. This action produces a stop pulse slightly longer than normal. The information from the read-only memory is loaded into the shift register by U5 during the time of the hard-wired start pulse.

When *reset/run* switch S3 is in *reset*, the 7493 counters are reset to zero, and the output shift register is allowed to run in a steady mark condition. When S3 is in the *run*, the counters begin to count the 64 different conditions for the address lines of the ROM with the 74121 one-shot loading the shift registers with each new output from the ROM. The first two outputs produce carriage returns, CR, and the final count, 63, produces a letter shift. The counter will return to zero and start a new line until the switch is flipped to *reset* or until the printer runs out of paper. (Actually, the printer won't stop, which could be rough on its platen).

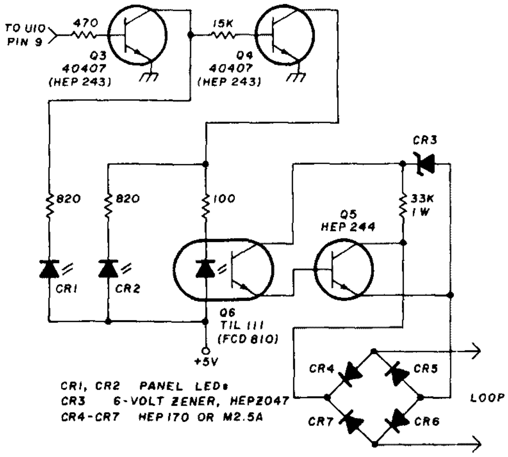


fig. 3. RTTY test-message generator output circuit, which is designed for standard 20- or 60-mil loops running at 100 volts minimum. Bridge voltage is high, so the M2.5A or equivalent diodes should be used.

output. The output circuit (fig. 3) is designed for standard 20- or 60-mA loops running at 100 volts minimum. It will not work with the so-called low-voltage loops. The diode bridge allows the loop to be connected with either polarity. Be sure to use diodes such as the M2.5A or equivalent, as the voltage will be high. Zener diode CR3 (5.6 volts) prevents the voltage from avalanching the opto-isolator. The opto-isolator can be an FCD-810, available from HAL.* The LEDs (optional) are for mark-space indication and are mounted on the control panel to indicate whether the system is in standby or run mode.

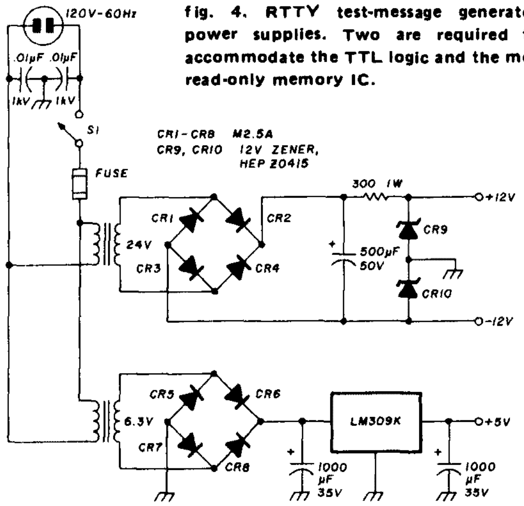


fig. 4. RTTY test-message generator power supplies. Two are required to accommodate the TTL logic and the mos read-only memory IC.

power supplies. Since both TTL and mos ICs are used, two power supplies are required; one for the TTL (plus 5V) and one for the mos logic (plus and minus 12V), fig. 4. For the 5-volt supply we used a 6.3-volt filament transformer. At the low-current drain of the TTL chips, the LM309K regulator works fine. However, it might be prudent to use 12 volts ac in the bridge for the 5-volt supply to make the LM309K work a little harder and provide somewhat better regulation.

The 12-volt supply, as shown, was used because it was simple. A center-tapped transformer could have been used but it would have required more components. The MM5220DF seems to be quite happy with the slight amount of ripple obtained with only 500 μF of filter. Use good-quality 12-volt zeners to ensure proper operation of the ROM. In both supply bridges, M2.5A 1kV 2.5A diodes were used because they were available.

*HAL Communications Corporation, Box 365, Urbana, Illinois 61801.

reference

1. Chet Opal, K3CUW, "The Micro-TO Keyer," *QST*, August, 1967, page 17.

ham radio

Since 1969, the year in which TEN-TEC introduced the first solid-state HF transceiver, progress has been rapid. It was in this year that we produced the POWER-MITE which triggered the booming QRPp activity. Two years later, the ARGONAUT followed, demonstrating that Ham Radio can be more than just push-button communication.

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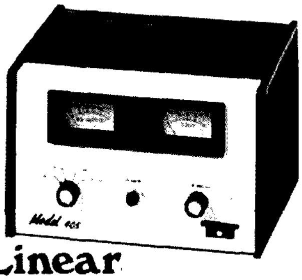
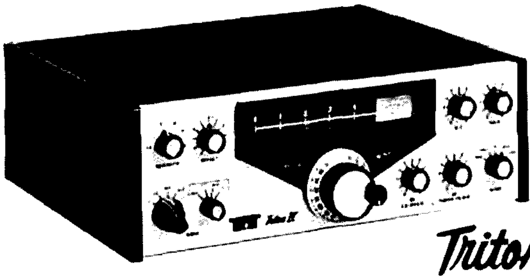


"405" Linear

100 watts input. Fully solid-state.



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More Details? CHECK—OFF Page 126

practical considerations in crystal-filter design

Construction and alignment techniques for building crystal filters that approach commercial standards

The principles of crystal-filter operation are well documented in the amateur literature; some excellent sources are found in references 1 through 3. In this article we'll explore the practical aspects of crystal filters, and show how they can be built with available crystals without resorting to sophisticated test equipment.

crystal selection

To the best of my knowledge only two sources of inexpensive high-frequency crystals are available: the surplus market and aviation equipment in which, until the late 1960s, crystals were used in large quantities. Frequency synthesizers have changed this situation, because today a single (or few) crystals control many different frequency channels.

The source of crystals is further narrowed because only those in type FT-243 holders should be used for the filters described here. These holders are sturdy and their construction permits convenient removal of the crystal for grinding. HC-6/U holders can't be easily opened for grinding in the home workshop. Crystals in HC-6/U holders can be used for the filters described here, however, by selecting those that will match the filter design requirements, which are discussed later. In any event, you should have a large quantity of crystals to choose from before beginning this project to ensure finding crystals that will match with satisfactory accuracy.

design example

Let's assume that a number of FT-243 crystals are available of equal nominal frequency. A crystal filter is to be built around these crystals. Whether the filter is to be used in a receiver i-f setup or in a transceiver is another matter. I recommend consulting references 4 and 5 in advance to make sure that the crystal-filter center frequency (which sets the i-f) will lead to a conversion scheme that will be as free as possible from spurious response.

The first step in selecting the four crystals used in the filter is to make sure they're as electrically similar to each other as possible. My standard procedure is to select crystals from the same manufacturer which are designed for the same parallel capacitance of the crystal-oscillator basic circuit. By so doing, I enhance the probability of ending up with crystals that are close relatives rather than distant cousins – a precaution that makes some of the later design steps easier.

activity check

Given the same oscillator circuit, some crystals have higher output than others. Your available crystals should be classified by grouping together those of approximately equal activity. This is easily done by inserting a meter in the collector circuit of the oscillator shown in **fig. 1**. Higher-activity crystals will show higher meter readings.

Next comes the frequency selection of the four crystals to be used in the filter circuit of **fig. 2**. From an electrical standpoint, the best procedure to find the pole-zero spacing is that reported in reference 2. However, the purpose of this article is to make matters as simple as possible, so we'll resort to another method.

crystal-frequency selection

To check the crystal resonant frequency using the circuit of **fig. 1**, the signal from the oscillator is injected into a frequency counter. If a counter isn't available, a communications receiver with a calibrated dial will do. What is recommended in the latter case is to read the crystal oscillator harmonics at as high a frequency as the receiver can cover. If, for example, the receiver goes to 30 MHz and the crystal fundamental frequency is 5.5 MHz, the harmonic at 5.5 x 5 = 27.5 MHz should be used. By so doing, the accuracy of the frequency readout is improved. Keep in mind that all we're concerned

By Jack Perolo, PY2PE1C, Caixa Postal 2390, Sao Paulo, Brazil

with is the *difference* in the resonating frequency of the oscillator-crystal combination when changing from one crystal to another. In other words, it is a secondary matter for this purpose if one crystal resonates at, say, 5,501.600 kHz or at 5,502.300 kHz; what really matters is to establish *exactly* how much higher or lower the resonating frequency is of crystal *A* vs *A'*; *B* vs *B'*.

Using this technique I have observed that, for a spacing of resonating frequencies of about 1.3 to 1.5 kHz, the resulting filter bandwidth (at -6 dB) is 1.8 to 2.1 kHz. Using a resonant-frequency spacing of 2.0 to 2.2 kHz, a suitable ssb filter can be obtained with a bandwidth at -6 dB of 2.5 to 2.7 kHz.

Referring to fig. 2, two matched pairs of crystals are needed for this type of filter. Using the technique just described, choose two crystals of like manufacturer, like nominal parallel capacitance (for the circuit), like activity, and like resonant frequency. The last requirement is the most difficult to achieve. I consider two crystals to be matched when, after meeting the first three requirements, they resonate in the same circuit with a *maximum* difference of 25 Hz. If this criterion can't be obtained with the crystals at hand, then one of the crystals must be ground, as outlined below.

If you're lucky, or have a large selection of crystals to start with, chances are that the four crystals needed for the filter can be obtained simply by proper selection without any grinding at all. If this seems possible, even though the filter bandwidth may end up slightly different from that required, I strongly recommend using the selected crystals to avoid grinding. Crystal grinding is an extremely delicate operation that's bound to cause some disappointment at the beginning.

The crystals must be matched pairs, so always start grinding the crystal of the planned pair that resonates at the *lower* frequency. The grinding operation *increases* the crystal resonant frequency.

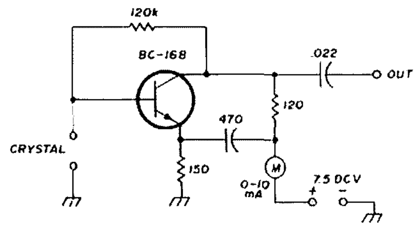
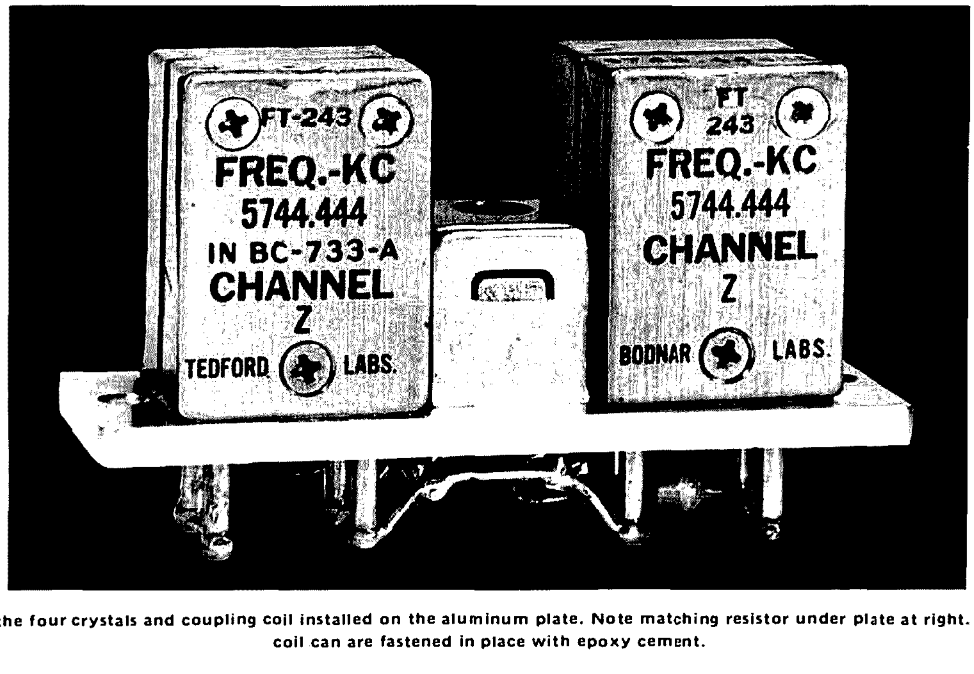


fig. 1. Oscillator circuit recommended for checking crystal activity and resonant frequency when choosing matched crystals for the filter.

Let's pick a numerical example, assuming the following crystals are available to build the filter:

A. 5501.267 kHz
B. 5501.291 kHz
C. 5502.018 kHz
D. 5502.326 kHz
E. 5502.120 kHz
F. 5501.914 kHz

Crystals **A** and **B** differ by only 24 Hz, so they need no further processing. If they were not matched, crystal **A** would be ground to move its resonant frequency closer to that of crystal **B**. If you're shooting for a difference in resonant frequency of 1.5 kHz for example, crystals **D** and **E** are recommended as a starting point. They should



Closeup of the four crystals and coupling coil installed on the aluminum plate. Note matching resistor under plate at right. Crystals and coil can are fastened in place with epoxy cement.

be ground to resonate at (5501.267 + 5501.291 kHz)/2 + 1.500 kHz = 5502.779 kHz ±12 Hz.

grinding procedure

Begin by preparing a scratch-free glass plate about 6 x 6 inches in area by about ¼-inch thick (152 x 152 x 6 mm). Wet the top surface of the glass with water and ground further. If its frequency is higher than this value, the crystal grinding was too extensive and this crystal is no longer useful for this particular filter. Select another crystal and repeat the process, reducing finger pressure and the number of strokes.

The procedure should be repeated until the crystal resonates at the desired frequency; then the second

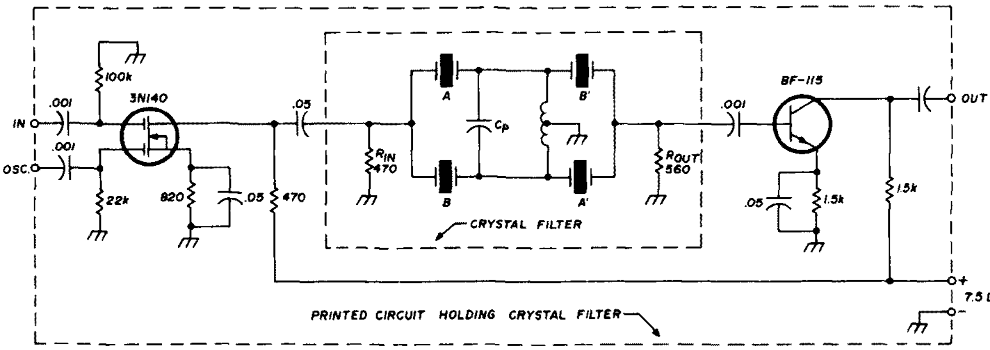


fig. 2. Schematic of the crystal filter using two matched sets of surplus crystals, together with input and output isolating stages. A pair of crystals are considered matched when they resonate in the same circuit with a maximum frequency difference of 25 Hz.

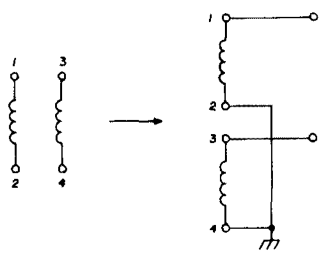
add a small amount of abrasive powder, 400 mesh or finer (available in hardware stores). Remove the crystal from its holder and, by *gently* applying pressure with one finger, move the crystal in a circular motion on the prepared surface of the glass plate.

It is extremely important that perfect parallelism be maintained between both sides of the crystal. After a few circular strokes, turn the crystal 90 degrees, apply the same number of circular strokes, and continue until the crystal has completed a full turn (360 degrees). Try to maintain the same finger pressure during each 90-degree segment of the grinding operation. Do not turn the crystal over during the grinding operation.

The ground crystal must now be washed in clean water and dried. Lay the crystal on a piece of absorbent material until all moisture evaporates. Next, clean the crystal again in a petroleum-derivative solvent; this operation ensures removal of any residual oil that may have been left on the crystal by your finger. Dry the crystal in open air and reinstall it in its original holder, taking pains to avoid touching the crystal surface with bare fingers. Use a pair of tweezers for this operation.

Check the crystal resonant frequency again, using the same original oscillator circuit and receiver setup. If its frequency is below 5502.779 kHz, the crystal must be

fig. 3. Bifilar-wound coupling coil for the two pairs of crystals used in the filter. For filters in the 5 to 6 MHz range, the coil consists of 7 + 7 turns of no. 28 AWG (0.3mm) enameled wire on a 10.7-MHz i-f transformer with a slug diameter of 3/32 inch (2.4mm).



crystal should be ground until its frequency matches its selected mate. In this process, more abrasive powder should be added if necessary. Two recommendations are in order at this point:

1. As the selected frequency approaches the desired value, a single stroke in the grinding operation can negate hours of work. The idea is to proceed slowly, avoid getting tired, and try to retain a parallel relationship between both sides of the crystal during grinding. You may have to remove and reinstall the crystal into its holder many times before satisfactory results are obtained.

2. If the crystal before grinding resonated at, say, 5502.390 kHz and after grinding its resonant frequency *decreased* to, say, 5501.920 kHz, then the crystal upper and lower surfaces have become out of parallel. You can either discard this crystal and start again from scratch, or if you want to recover the crystal, its thickness should be measured and the crystal should be ground until parallelism has been restored, bearing in mind that grinding causes the crystal frequency to increase.

Grinding a crystal out of parallelism to decrease its frequency is not recommended, as crystal activity will be decreased. Furthermore, the chances of obtaining spurious response from the filter are increased.

filter assembly

Matching the four crystals in two pairs is the most time-consuming part of the project. The crystals must be assembled on a subchassis together with the coupling coil. I use an aluminum plate measuring 1 x 3 x 1/8 inches (25 x 76 x 3 mm). The crystals and coil are fastened in place with epoxy cement. The coil must be bifilar wound, with as close coupling as possible. If the

coils are wound simultaneously, the connections shown in fig. 3 should be made to obtain the desired bifilar configuration. To ensure close coupling, the coil should have a slug of the "closed cup" type, or a toroid should be used. Tuning a toroid is more critical, and that's why I prefer the closed-cup variety.

filter alignment

Different precautions must absolutely be taken if the filter is to perform comparable to commercial units. The crystals must have been thoroughly cleaned after the grinding procedure, or else the filter will perform erratically as a function of time.

The input and output leads should be as short and rigid as possible a change in length of the input and/or output leads will cause the filter to have a dramatically different bandpass response. To avoid this problem, I install the filter over a printed circuit board with isolating stages before and after the filter. By so doing, I'm assured that filter alignment will remain steady over the years and remain independent of filter application.

Whatever convenient setup is used, filter alignment can be made using different approaches. Evidently the best is to use a signal generator and a scope with a sweep system so that filter response may be directly displayed on the scope and adjusted accordingly. Since this is a rather sophisticated technique requiring some expensive test equipment, I assume that those who have a scope also know how to use it; consequently no description of this option is given. Instead I'll describe an easier alignment procedure, although it's more time consuming.

The setup of fig. 4A is recommended for the filter alignment. If no vfo is available with the required frequency and stability output, a suitable alternative method is shown in fig. 4B.

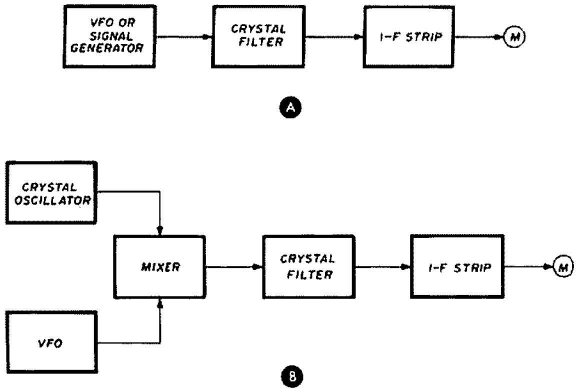
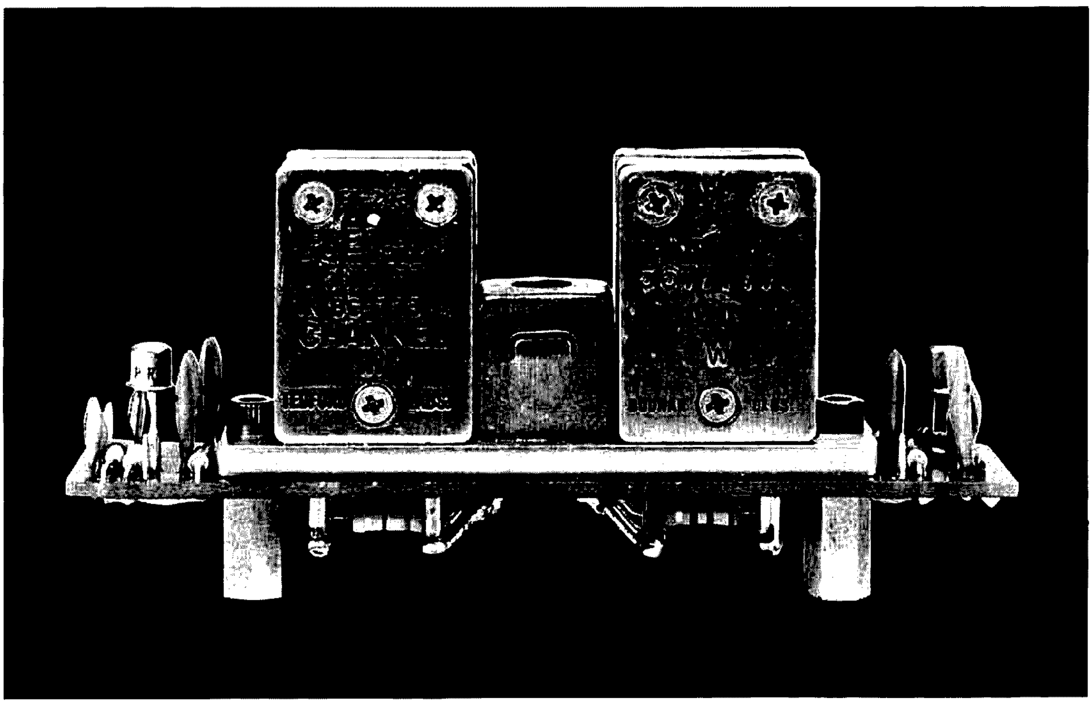


fig. 4. Setup for filter alignment. The setup at A is the recommended arrangement; an alternative method is shown in B. In both cases the vfo should be stable and accurately calibrated.

The approximate center frequency of the filter should be determined, then filter response should be determined by varying the vfo frequency ±3.0 kHz around the center frequency. This operation should be performed at 100- or 200-kHz increments, recording the vfo reading and the corresponding meter reading on paper in each case. If a calibrated vfo isn't available, its frequency can be determined with good accuracy by using the technique previously described for determining the frequencies of the crystals.



Crystal filter installed on a printed circuit board, complete with an isolating stage ahead of i-f (with a 3N140 transistor) and another following it (with a BF115 transistor). Pillars support the whole assembly.

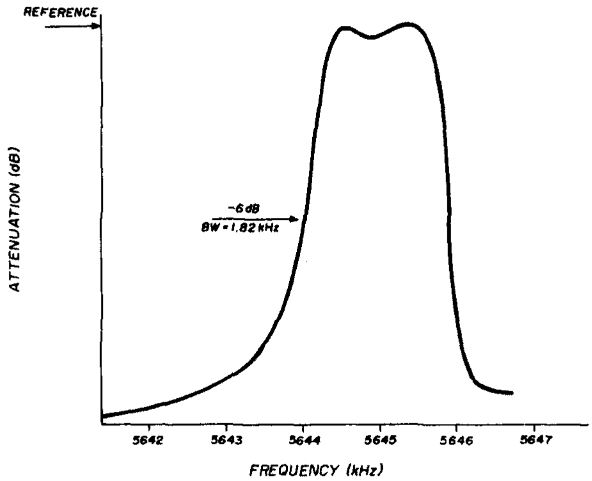


fig. 5. Bandpass characteristics of a 5645-kHz four-crystal filter built by the author using the technique described in the text. Frequency differences in the two matched pairs were 6 and 14 Hz, respectively. Insertion loss of this filter was about 5 dB.

At this point four variables are at hand to achieve a) the least amount of insertion loss, and b) the flattest bandpass response:

1. Value of R_{in}.
2. Value of C_p.
3. Position of the coil slug.
4. Value of R_{out}.

While the value of R_{in} and R_{out} can't be too far from 560 ohms or so, items 2 and 3 may vary considerably from one case to the next. For filters in the 5 to 6 MHz range, I've consistently obtained good results using 7 + 7 turns of no. 28 AWG (0.3mm) enameled wire for the coil. The coil form is a 10.7-MHz i-f transformer with a slug diameter of 3/32 inch (2.4mm). The value of C_p will vary between 27 and 68 pF; 39 to 47 pF is the most common value.

Fig. 5 shows the bandpass characteristics of a 5645-kHz crystal filter built recently using the technique described. The coil used in this filter has 7 + 7 turns, as described; C_p = 39 pF, R_{in} = 470 ohms, and R_{out} = 560 ohms. The passband ripple is below 1 dB and the 60/6 dB shape factor is about 1:2.1. The crystals were originally for 20 pF parallel capacitance and were ground and checked with the oscillator of fig. 1. They were considered to be matched when they displayed these operating frequencies:

$$\left.\begin{array}{ll} A & 5644.410 \text{ kHz} \\ A' & 5644.416 \text{ kHz} \end{array}\right\} \Delta f = 6 \text{ Hz}$$

$$\left.\begin{array}{ll} B & 5645.627 \text{ kHz} \\ B' & 5645.641 \text{ kHz} \end{array}\right\} \Delta f = 14 \text{ Hz}$$

The measured bandpass at −6 dB is 1.82 kHz; two spurious responses were recorded 13.3 and 18.3 kHz above the bandpass center frequency. The insertion loss of the filter is about 5 dB.

Several distorted patterns (all of which were obtained at one time or another while the filter was being calibrated) are shown in fig. 6; all indicate that some adjustment was missing on the filter. It goes without saying that a system with four variables deserves some respect and, unless the alignment problems are properly tackled, no result will be achieved. Consequently only one parameter at a time should be varied to obtain useful conclusions.

ripple, spurious response, and insertion loss

Passband ripple can be adjusted to as low as 0.3 dB, an extremely good figure even by commercial standards. Spurious response generally appears as one or two signals some 10 or 20 kHz removed from the filter center frequency. Both responses generally show the same attenuation, about 40-35 dB below the midband signal.

Insertion loss becomes generally higher by attempting to obtain a ripple-free, perfectly symmetrical bandpass response. At any rate, I do not consider the insertion loss a critical item, as this attenuation can be easily compensated by adding an extra stage of amplification after the filter, whereas other characteristics of the crystal filter cannot be externally compensated.

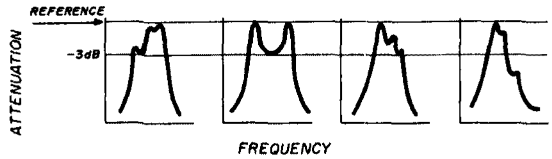


fig. 6. Examples of bandpass response of improperly adjusted filters, indicating that some important step or steps were missing when calibrating or grinding the crystals or when alignment was attempted.

It must be remembered, however, that the filter response may be substantially deteriorated if the i-f strip following the filter is misaligned; I recommend checking the entire i-f response, complete with the filter, to make sure that nothing has gotten out of hand. No shielding of the filter, as shown in the photos, is necessary as long as it is operated away from high-powered stages.

acknowledgement

Thanks go to Maiso, PY2GP, for many of the observations reported and for much guidance on filter construction.

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ham radio

transmission line calculations

with your pocket calculator

How to use your four-function calculator to design transmission-line matching transformers and matching stubs

As many amateurs have discovered, cut and try transmission-line adjustments can become a frustrating experience. Strange things have been known to happen when you fiddle around with different lengths of coaxial or open-wire line. When you dig out a textbook on transmission-line theory, those hyperbolic functions and associated complexities can be intimidating, to say the least!

In recent years the Smith chart has been *the* engineer's tool, and those of us who were able to meet it on even terms acquired a new and different concept of transmission lines. Unfortunately, Smith charts are not the easiest things for the amateur to acquire, and although its application and use have been described in the amateur magazines,[1,2] its mastery takes a good deal of practical use.

More recently, the computer has appeared on the scene, but few of us have access to its mysteries, so once again the average amateur is left out in the cold. However, a ray of hope has appeared in the shape of the popular hand-held electronic calculator. If you will compromise to the extent of a book of trigonometric tables, the humble four-function calculator can do a pretty good job of coming up with the answers. Where do you locate that matching stub or transformer? If your handy-dandy device is sophisticated enough to do trig functions as well, you won't even need the trig tables!

To begin at the beginning, consider a length of transmission line, of characteristic impedance Z_o, terminated by a mismatched load having a resistive component, R, and a reactive component, X (see fig. **1**). If you now refer to the Smith chart of fig. **2**, you will see that it has a scale around its circumference which is labeled "Angle of Reflection Coefficient in Degrees." You will also see that a set of rectangular coordinates has been superimposed on the Smith chart. The locus of all the centers of reactance circles has become the X axis, while all of the centers of resistive circles are located on the Y axis. This allows us to express Smith chart functions in simple trigonometric terms, and eliminates the complex j operator, which the simple pocket calculator cannot handle.

This article will show how, given a complex load of $R + jX$ (or $G + jB$), appropriate points may be computed where either a suitable matching transformer or shunt stubs may be located to match out the transmission line.

By Henry S. Keen, W5TRS, Fox, Arkansas 72051

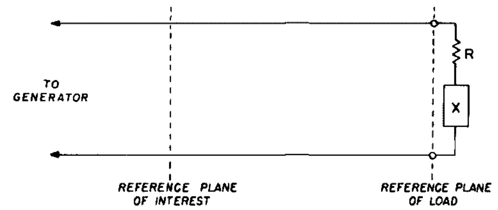

fig. 1. Transmission line terminated with a complex load. Design of matching transformers or matching stubs, and their proper location, may be determined with a Smith chart, or by using the simple four-function calculator, as discussed in the text.

In addition, a general relationship will be given, with which the impedance (or admittance) resulting from looking at a known complex load through a given length of transmission line may be determined.

It may also be determined from the Smith chart that as you back off from the load, moving toward the generator (transmitter), that the angle of the reflection coefficient will be *reduced* as you recede toward the generator.

There are two general methods of matching a mismatched transmission line to eliminate standing waves. The first is to insert a matching transformer (usually a quarter wavelength of line, of a different characteristic impedance than the line to be matched). The second, which is more adaptable to coaxial lines, is to install a shunt stub at the proper spot on the line.

The location for a matching transformer, if you are using open-wire lines, is quite simple as you only have to run a neon lamp along the line to locate a voltage minimum or maximum. These are points where the reactive component is zero. If plotted on a Smith chart, this impedance would fall upon the "real" diameter, which is marked as the Y axis in **fig. 2**. At this point you would cut the line and insert a matching transformer having a transformation ratio equal to the square root of the vswr. Whether its characteristic impedance would be higher or lower than the main transmission line depends upon whether you had chosen a minimum or a maximum voltage point.

With coaxial lines, however, you cannot detect voltage maximums and minimums as you move a neon lamp along the line, so all points have to be computed, based upon knowledge of the complex impedance of the load. Bridges for measuring such complex quantities have been fairly well covered in the literature, and need not be discussed here.[3,4]

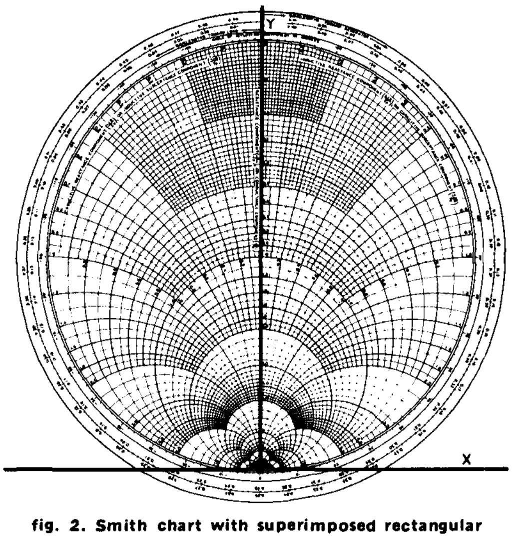

fig. 2. Smith chart with superimposed rectangular coordinates (see text).

transformer location

Starting with the load impedance of $R_o + jX_o$ you must first normalize the load by dividing both the resistance and reactance of the load by the characteristic impedance of the line. Thus, $30 + j60$, normalized to 50 ohms, for example, would be $30/50 + j60/50 = 0.6 + j1.2$.

Secondly, you must find the angle of the reflection coefficient, θ, using the relationship

$$\tan\theta = \frac{2X}{R^2 + X^2 - 1} \tag{1}$$

In this usage, whenever X is positive, θ also is positive, and when X is negative, θ is negative, as on the Smith chart. When the denominator is positive, θ lies between zero and 90°, and when the denominator is negative, θ lies between 90° and 180°.

In the case of our assumed impedance of $0.6 + j1.2$, for example;

$$\tan\theta = \frac{2(1.2)}{0.36 + 1.44 - 1} = 3$$

Therefore $\qquad \theta = 71°34'$

Inasmuch as the length of line included affects both the incident and the reflected wave, the transformer should be located 35°47′ (half of 71°34′) back from the load. Such accuracy as represented by the 47′ is unnecessary for any amateur operation, but it's nice to know that you can figure it that closely (assuming, of course, that you accurately measure the load impedance). This is a point of low impedance (0°). If you wanted to locate the transformer at a high impedance point, it would be located an additional 90° back from the load.

The vswr represented by this load may be calculated by first finding the reflection coefficient, Γ from the following expression:

$$\Gamma = \frac{2X}{[(1+R)^2 + X^2]\sin\theta} \tag{2}$$

Solving **eq. 2** for the example above

$$\Gamma = \frac{2(1.2)}{[(1+0.6)^2 + 1.2^2]\sin 71°34'}$$

$$= 0.63245$$

The corresponding vswr is

$$vswr = \frac{1 + 0.63425}{1 - 0.63425} = 4.44$$

Therefore, the transformation ratio would be $\sqrt{4.44}$ or about *2:1*.

When coaxial transmission line is used, you rarely have the chance to select the characteristic impedance of a matching transformer, so the matching must usually be accomplished by means of shunt stubs, which can be calculated in the following manner: Convert the complex admittance $R_o + jX_o$ to the admittance form, as follows, using the normalized resistance and reactance

$$Conductance,\ G_o = \frac{R_o}{R_o^2 + X_o^2} \quad (3)$$

$$Susceptance,\ B_o = \frac{X_o}{R_o^2 + X_o^2} \quad (4)$$

In transferring from impedance to admittance, the sign changes, so that $R_o + jX_o$ becomes $G_o - jB_o$ (and $R_o - jX_o$ becomes $G_o + jB_o$). This is because inductance is considered a positive reactance, while inductive susceptance is negative.

stub matching

1. Using the admittance counterpart of **eq. 1** find θ from the expression

$$tan\ \theta = \frac{2B_o}{G_o^2 + B_o^2 - 1} \quad (5)$$

2. Using the admittance counterpart of **eq. 2**, find Γ from the expression

$$\Gamma = \frac{2B_o}{[(1 + G_o)^2 + B_o^2]\ sin\ \Phi} \quad (6)$$

3. The admittance must move down the transmission line to a point where $G_1 = 1.0$, at which point the new value for B may be found from

$$B_1 = \pm \frac{2\Gamma}{\sqrt{1 - \Gamma^2}} \quad (7)$$

Substitution of this new value of susceptance, B_1, with $G=1$ in **eq. 5** will give the value of θ at the location for the stub. If B_1 should be positive, the added stub will be negative (inductive). You can choose either type of stub, but the widest bandwidth will be obtained when you choose the point closest to the load.

As an example, let's return to our original assumed load impedance which normalized to $0.6 + j1.2$.

From **eq. 3** $G_o = \frac{0.6}{0.6^2 + 1.2^2} = \frac{0.6}{1.8}$

$$= 0.333$$

$$B_o = \frac{1.2}{0.6^2 + 1.2^2} = \frac{1.2}{1.8}$$

$$= 0.667$$

Therefore, the admittance equivalent is 0.333 - j0.667

From **eq. 5** $tan\ \theta_o = \frac{2(-0.667)}{0.333^2 + 0.667^2 - 1}$

$$= \frac{-1.333}{-0.444} = 3.000$$

from which we deduce that θ_o is negative (B_o is negative) and that θ_o is between $-90°$ and $-180°$ (denominator is negative).

$$arctan\ 3.000 = 71°34'$$

The value of θ_o is therefore $-180° + 71°34' = -108°26'$

The reflection coefficient, Γ, has not changed from previous calculations and is still 0.63245. When this value is used in **eq. 7**

$$B_1 = \pm \frac{2(0.63245)}{\sqrt{1 - 0.63245^2}}$$

$$= \frac{1.2649}{\sqrt{0.600}} = 1.633$$

Therefore, the nearest point to the load, where a shunt stub could be located, is where $Y_1 = 1.0 + j1.633$. Consider clockwise rotation on the Smith chart from $-108°26'$, so applying **eq. 5** again, using these figures,

$$tan\ \theta = \frac{2(1.633)}{(1.0)^2 + (1.633)^2 - 1}$$

$$= 1.2247$$

$$\theta = 50°46'$$

Therefore, the total distance will be $(180° - 108°26') + (180° - 50°46') = 200°48'$ This will require a length of line half that, or $100°24'$ *(0.2789 wavelength)*.

Since B_1 is positive, the stub must be inductive and present a shunt susceptance equal to -1.633. The stub length is determined from

$$cot^{-1}\ 1.633 = 31°29'\ (0.08745\ wavelength)$$

The characteristic impedance of the stub is assumed to be the same as the main transmission line.

If you had wanted to use a capacitive stub (or a shunt capacitor), you'd have chosen B_1 to be -1.633, which would have placed the new value of θ_1 as $-50°46'$, and would have called for a length of 0.4247 wavelength between the load and the shunt susceptance.

general relationship

When a given normalized complex admittance, Y_L, is to be translated a distance θ (in electrical degrees) down a line of admittance Y_o, the classical formula is

$$Y_1 = \frac{Y_L + j\ tan\ \theta}{1 + jY_L\ tan\ \theta} \quad (8)$$

This relationship looks innocent enough, but when Y_L is complex to begin with, then you must, after expanding and combining terms, also rationalize the denominator. There are numerous pitfalls which can trip up the unwary (when I used this expression, I always checked the results on a Smith chart just to be sure).

However, the same translation of a complex admittance can be accomplished in the following manner:

Given: $Y_L = G_L + jB_L$ (or $Z_L = R_L + jX_L$)

to extend the transmission line by θ degrees and find Y_1 (or X_1)

1. From **eq. 5**, find θ_o (or use **eq. 1** for Z)
2. From **eq. 6** or **eq. 2** find Γ
3. Subtract the quantity 2θ from θ_o, giving θ_1
4. Find $X_1 \doteq \Gamma \sin \theta_1$ and $Y_1 = 1 - \Gamma \cos \theta_1$ **(9)**
5. Calculate

$$G_1 = \frac{2Y_1}{X_1^2 + Y_1^2} - 1 \quad (\text{or } R_1) \qquad \textbf{(10)}$$

$$B_1 = \frac{(1 + G_1)X_1}{Y_1} \quad (\text{or } X_1) \qquad \textbf{(11)}$$

substitute R_1 for G_1

I feel that this process is not as messy as **eq. 8**, in addition to being compatible with the pocket calculator, particularly those which have trigonometric capability. It may be that some or all of these steps might be programmed on the more sophisticated HP-25 or HP-65 calculator, which would make the process a breeze.

Another application of these relationships might be to translate the impedance of an antenna, as measured through a length of coaxial cable, *back* up to the antenna itself. In such an application, the length of the coaxial line in electrical degrees must be accurately known for each frequency of measurement.

When moving *toward* the load, instead of away from it, as was done in the foregoing discussion, twice the electrical length of the line through which measurements are taken must be *added* to the calculated angle of the reflection coefficient. Integral half wavelengths (180°) of line, which result in a complete revolution on the Smith chart, would of course be discarded. Therefore, if you neglect losses in the transmission line, the electrical characteristics of the antenna might be determined to a fair degree of accuracy without disturbing the electromagnetic field around it. The use of a carefully measured and known length of coaxial line for this purpose would be most helpful. A standard test line should be of considerable value to the antenna experimenter as a primary piece of test equipment.

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ham radio

repair bench

Hank Olson, W6GXN

power-supply servicing

Probably the **most** common **service job** in electronic equipment involves the power supply. The power supply in amateur equipment is usually connected to the 110- or 220-volt ac line through switches, fuses, or circuit breakers. This connection is a significant interface between your equipment and the real (and imperfect) world. If your power supply is not connected to a source of perfect sine-wave power, then you can expect problems, especially if the supply is used to power solid-state equipment. In this month's column we discuss a number of power supplies and voltage regulators and how to troubleshoot problems that may develop – what to look for, how to isolate faults, and how to make repairs.

line-voltage transients

The most common power-supply problems don't occur from plugging the supply into a source of the wrong line voltage; such errors usually result in an open primary fuse or open circuit breaker, which is easily fixed. A more important problem occurs from line-originated transients, which cause the input voltage to depart from a pure sine wave. A line-voltage transient can be a positive high-voltage spike riding on the most-positive excursion of the 60-Hz sine wave (fig. 1). The duration of this spike is very short, so it doesn't contain enough energy to open a fuse or trip a circuit breaker. However, it can appear on the power-transformer secondary winding (if the transformer has sufficiently high frequency response) and result in overvoltage on one or more of the rectifiers.

Overvoltage occurs when the rectifier peak inverse voltage (PIV) exceeds the rectifier rating. In tube rectifiers this is called flashback; in modern silicon rectifiers it's usually called *destruction*! Some silicon rectifiers (more expensive types) are made to recover from low-duty-cycle overvoltage; these are the "avalanche-protected" variety. Because they're more expensive, you probably won't find them in your commercial amateur equipment. Some of the principal makers of these avalanche-protected diodes are Semcor, Unitrode, and Varo. What you're most likely to find in your commercial gear is the 1N4001-1N4007 series or something similar, which are *not* protected and which are worth about 5 cents each in OEM quantities. If they fail, you may not find these rectifiers in your equipment except for their wire ends; the package will have disintegrated into small black granules rattling around in the bottom of the equipment case.

If the above seems overstated, it was because I wanted to make a point: by far, *the most common fault in power supplies is rectifier failure.* In modern circuitry using silicon rectifiers, the diodes will most often be shorted; the only open-circuited rectifiers I've seen are those resulting from package fracture or disintegration.

typical circuits

To troubleshoot power supplies, we must first know what circuits might be encountered. In **fig. 2** are seven common rectifier-filter configurations, all using capacitor-input filters. The first is the simple half-wave rectifier, which is the least expensive, but which produces 60-Hz ripple (making filtering more difficult). The second is the full-wave rectifier, which requires a center-tapped transformer. The third is the full-wave bridge, which requires no center tap but which requires four rectifiers. The conventional voltage doubler is shown next, followed by the cascade voltage doubler. The full-wave, full-wave bridge, and conventional voltage doubler all produce dc output with 120-Hz ripple. The cascade voltage doubler produces 60-Hz ripple very much like the simple half-wave circuit.

The last two circuits of **fig. 2** are variations on the full-wave bridge. The circuit in **fig. 2F** produces both positive and negative voltages (as in power supplies used

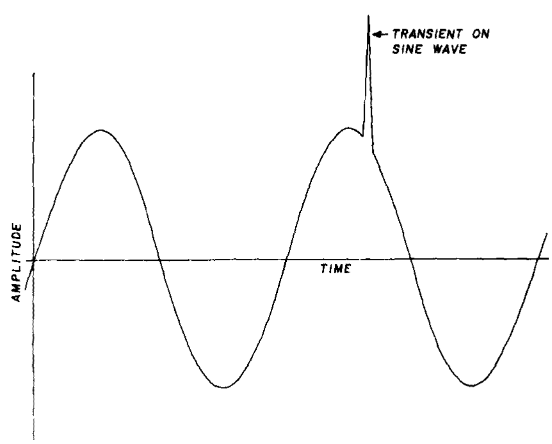
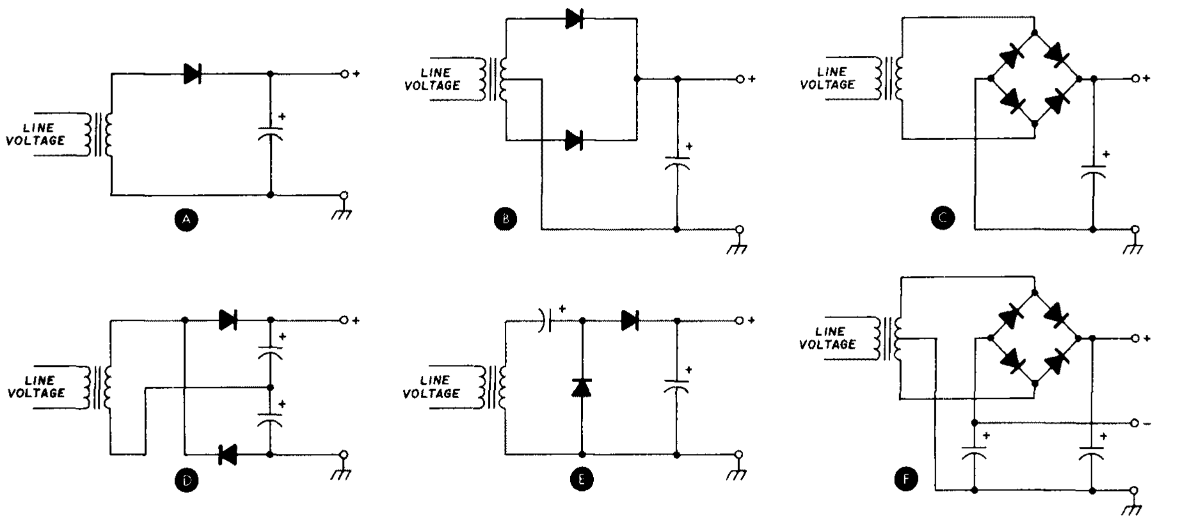


fig. 1. Representation of a 60-Hz line-voltage wavetrain with a transient spike, which can cause overvoltage on power-supply rectifiers.

By Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

for operational amplifiers). Looking at this circuit carefully, we see that it boils down to simply a positive and a negative full-wave rectifier operating from the same center-tapped transformer. The other full-wave bridge, fig. **2G**, is the dual-voltage version. It produces *half* the dc output at the transformer center tap as produced at

Power supplies that do have a central regulator, however, generally use some sort of series circuit, much like those in fig. **4**. Figs. **4A** and **4B** show the positive and negative versions. The negative regulator works in exactly the same way as the positive regulator and will be more common in older designs that used germanium



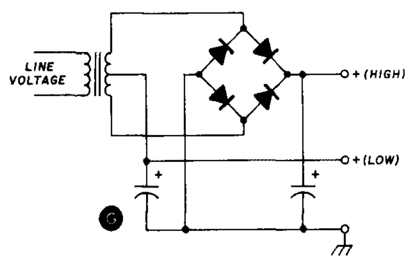


fig. 2. Typical power-supply circuits. A, B, and C depict half-wave, full-wave, and bridge rectifiers using capacitive-input filters. A conventional voltage doubler in D and a cascade voltage doubler in E. Variations of the bridge are shown in F and G, the former producing both positive and negative voltage and the latter producing half the dc output at the +low terminal as produced at the +high terminal.

the plus terminal of the bridge. This supply is similar to that of fig. **2F** but with a different point grounded.

Seven rectifier-filter circuits with typical component values, together with plots of voltage and current output, are shown in fig. **3**. The graph of fig. **3H** shows that wide variations occur in the output of the different circuits even though the same transformer and filter-capacitor values are used. Although choke-input filtering is not too common in modern commercially built amateur gear, it's possible to use such filtering with the full-wave and full-wave bridge rectifiers. Typical examples are shown in figs. **3F** and **3G**.

voltage regulators

Following the rectifier-filter section there will frequently be no signal regulator, because many circuits in commercial electronic equipment are not very sensitive to voltage variations. In a typical receiver, for instance, the only stage that may use regulated +Vcc is the local oscillator. In even the best and most expensive designs, the audio output stages are operated from unregulated +Vcc.

power transistors. The only differences between the designs are that, in troubleshooting the positive regulator, one measures positive voltages to common; in the negative regulator, one measures negative voltages to common. The other major difference, when fixing older germanium designs, is that the base-emitter forward voltage is closer to 0.3 volts whereas it's about 0.6 volt for silicon transistor designs.

Perhaps the simplest series regulator design is the emitter-follower, shown in fig. **5A**. In this case no separate feedback of the output voltage to the control section is used, but it is assumed that the base-emitter voltage is more or less constant. The base is held at constant voltage by the zener diode, and ripple at the base is reduced by the filter formed by R1, C1. The next most-complex series regulator is shown in fig. **5B**, where feedback is used from the output. A fraction of the output voltage is fed to Q2 base. The difference between this voltage and the zener diode voltage (plus the emitter-base forward drop of Q2) is amplified to control Q1 base. The third most-complex regulator uses a differential amplifier as a control circuit with one side refer-

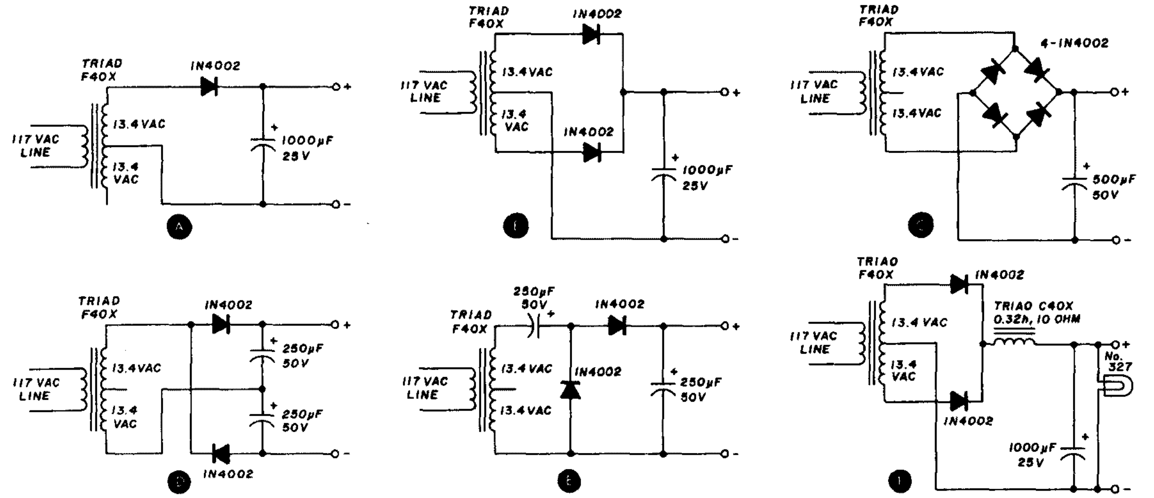
enced to a zener and the other to a fraction of the output voltage. This circuit is shown in fig. 5C; note that a second zener is usually used to provide coarse regulated voltage to the differential pair.

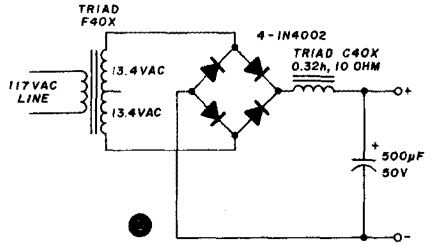
IC op amp regulators

At this point it is advantageous to substitute an operational amplifier for the differential pair. Many

Q1 base in a negative direction. This action prevents further current being passed by Q1. If Q2 were a germanium transistor R2 would have to be about 3 ohms to allow current limiting at 100 mA, again because of the lower base-emitter forward drop of germanium transistors.

In the next generation of linear ICs the voltage-reference and current-limiting functions and even a small





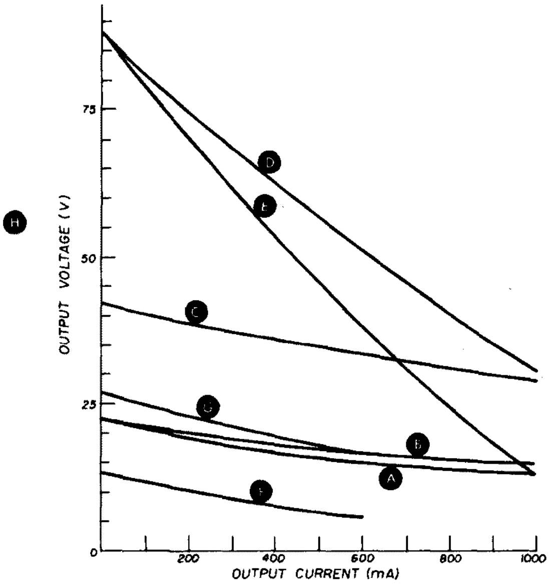


fig. 3. Rectifier-filter circuits with typical component values. A through E are respectively a half-wave rectifier, full-wave rectifier, bridge rectifier, voltage doubler, and cascade voltage doubler. Full-wave and bridge rectifiers are shown in F and G, using choke-input filters. Response curves of these systems are shown in H. The curves for the choke-input systems were terminated at 600 mA because of the choke current rating.

commercial regulators that were designed shortly after monolithic IC amplifiers became available are of the form shown in fig. 6.

In fig. 7 are actual circuits of the regulators discussed above. Note that in fig. 7D a current limiter has been added consisting of Q2, R1, and R2. (Compare with fig. 6). This feature can be added to most regulators and is certainly worthwhile. When 100 mA is drawn, 0.6 volt is developed across R2, causing Q2 to conduct, which pulls

series-pass transistor were integrated into the regulator chip. National's LM300, Fairchild's µA723, and Motorola's MC1460 are representative of this first generation of IC regulators. In their small TO-5 packages these regulators could handle currents of only 10 mA or so, depending on the input-output voltage and the adequacy of the heat sink. These ICs can be used with external power transistors to provide regulators with *amperes* of current capability. Fig. 8 shows circuits using these ICs

with external transistors as combined regulators. The LM300 has since been replaced by the improved version, LM305, and the MC1460 has also been replaced by the MC1469. The μA723 and LM305 are widely second-sourced, so you may find them with any number of brand names.

Finally, in the second generation of IC regulators, manufacturers have succeeded in putting even the power transistors onto the chip. Such regulators are typified by the Fairchild μA7800 and National LM340 series of three terminal regulators, in power packages. These ICs have fixed output voltages; you simply buy the voltage type desired. The negative three-terminal regulators, μA7900 and LM320 series, operate in similar fashion. Of course, since there is no user control of output voltage or current limiting on any particular three-terminal regulator, there is also nothing to repair once the input, output, and ground connections are checked for proper connections and voltages; IC replacement is the only remaining option. The use of three-terminal regulator ICs is shown in fig. 9. Note that a small capacitor is necessary between input and common for stability.

troubleshooting and repair

Preliminary checks. Now that we've covered at least a fair number of rectifier-filter and regulator circuits most likely to be encountered, let's move on to fixing them. It is perhaps superfluous advice, but your first check should be to see if line voltage is actually entering the equipment under repair. Line cords, especially the molded variety, are frequently open (and occasionally not plugged in). Next, fuses and breakers should be checked; even if they appear OK, check them with an ohmmeter. Apparently good fuses can occasionally open in a way that will be missed by a visual check. If new fuses immediately blow, then you must resist the temptation to use much-larger-than-normal fuses to "get things going." Such overriding of the fuse function, even for "a short time during servicing," will usually cause severe damage to some component that may be hard to replace — such as a specially made power transformer.

Rectifiers. The next step is to check the rectifier diodes using an ohmmeter, as mentioned before, these are the most likely candidates for failure. It is easiest to disconnect one lead of each rectifier before applying the ohmmeter test, otherwise the dc resistance of the power

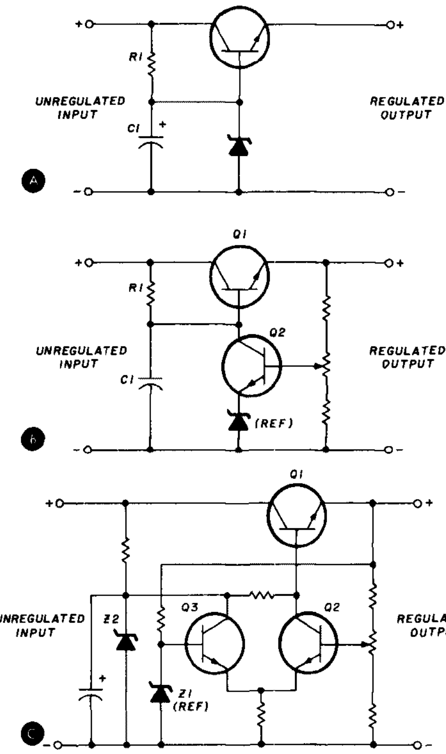


fig. 5. Simple emitter-follower series regulator, A, in which base-emitter voltage is held constant by a zener. In B, a fraction of the output voltage is fed back to Q2, a dc amplifier, to control the base of Q1. The series regulator in C uses a differential amplifier, Q2, Q3, as part of the control circuit.

transformer secondary can confuse the issue. If a bad rectifier is found, replace it with a new part or equivalent. The Motorola "HEP Cross Reference Guide and Catalog" is the best of the several replacement handbooks in this area, in my experience. My advice is to replace all the diodes in a full-wave or a full-wave bridge rectifier (unless you have an *exact* replacement) if only one rectifier diode is shorted. In this way, the original balance of diodes is maintained. The required PIV in the replacement diodes should be twice that of the originals to ensure reasonable reliability.

Such diodes are very inexpensive. As an example, consider a full-wave rectifier with capacitor input and an output voltage (unloaded) of 15 volts. You could get away with using a nominal 50-volt PIV diode such as the 1N4001 and still have a 10-percent margin of PIV before breakdown. A 1N4001 costs 32 cents (1976 Allied

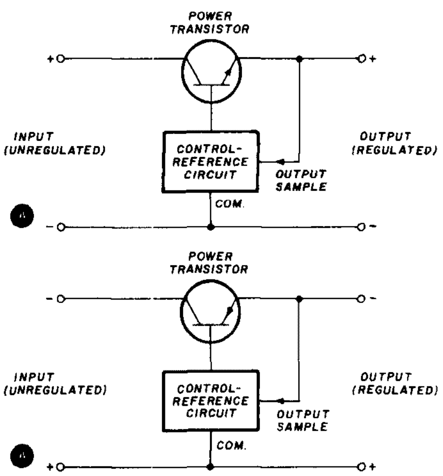


fig. 4. Series regulators. Positive and negative versions are shown in A and B. The negative regulator is more common in older designs using germanium power transistors.

Radio Catalog), and a 1N4002 costs 33 cents. The difference is *one cent* in price, but this one cent buys you *double* the PIV (100 volts). It certainly is a worthwhile penny spent! Diodes (up to about 15 amps) are so cheap because of wide consumer use – it's ridiculous to pinch pennies on PIV.

Transient suppressors. After replacing rectifier diodes in a power supply, should a repetition of the failure occur in a few weeks or months, it may be wise to add something to suppress line transients. A 0.01-μF 1kV disc capacitor across the transformer secondary will do a lot to suppress short-duration spikes for example. An even more effective spike reducer is a thyrite varistor of the correct voltage. Disc and rod varistors are rated in voltage (for both ac and dc) by their manufacturers. These components may also be connected across the transformer secondary. Such thyrite varistors are made by General Electric, National Lead, and Automatic Electric.

The varistors have the advantage of *dissipating* the transient as opposed to the capacitor, which integrates the transient energy into a lower voltage, longer impulse. In power supplies using choke-input filters, a thyrite varistor may be used as in fig. 10 to prevent the choke-field collapse voltage from exceeding the rectifier diode PIV when the power supply is turned off.

Load circuitry. After the rectifiers have been given a clean bill of health, disconnect the regulator output from the circuit that it is designed to power. A failure in the circuitry that loads the regulator can give the appearance of a regulator failure, especially if the regulator has built-in current limiting. If low output voltage still occurs, the regulator input voltage should be checked. If the regulator input voltage is low, disconnect the rectifier-filter output from the regulator input. This allows the rectifier-filter to operate unloaded, and its output voltage should increase at least to nominal voltage when turned on. We have now three different system blocks where an apparent power-supply problem can occur: rectifier-filter, regulator, and load circuitry. By disconnecting these blocks from each other in steps, the faulty section can be isolated and fixed.

filter capacitors

If the trouble is in the rectifier-filter section, the most

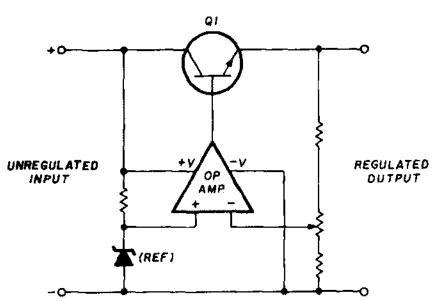


fig. 6. Series regulator using an op amp IC. Circuit is typical of many commercial designs that appeared after monolithic IC op amps became available.

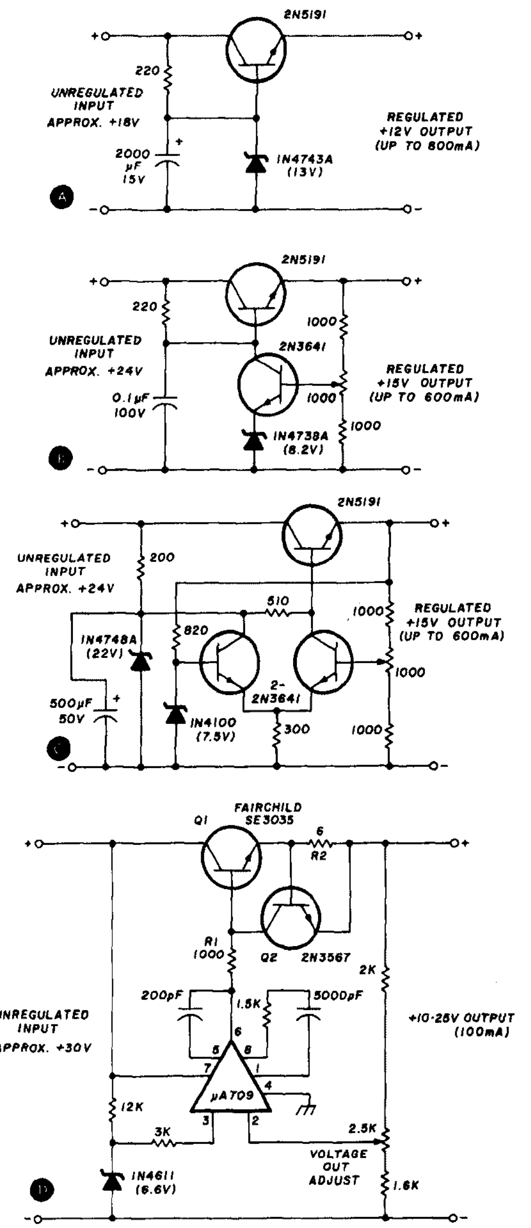


fig. 7. Series regulator circuits of figs. 5 and 6 using typical components. The emitter-follower version is shown in A; simple series regulator, B; series regulator using differential dc amplifier, C; and a series regulator using an IC op amp as a differential amplifier, D. The current limiter consisting of Q2, R1, R2 in D is a simple and worthwhile addition that can be included in most regulators.

probable failure (after rectifier diodes) is an electrolytic filter capacitor. The failure of electrolytic capacitors can either be a short or open circuit. A shorted electrolytic can be easily found with an ohmmeter; the only indication of an open electrolytic is its failure to smooth out ripple. Such an open-circuit failure will cause the dc voltage to decrease to a value lower than the peak ac voltage, as read on most meters.

An oscilloscope readily shows an open filter capaci-

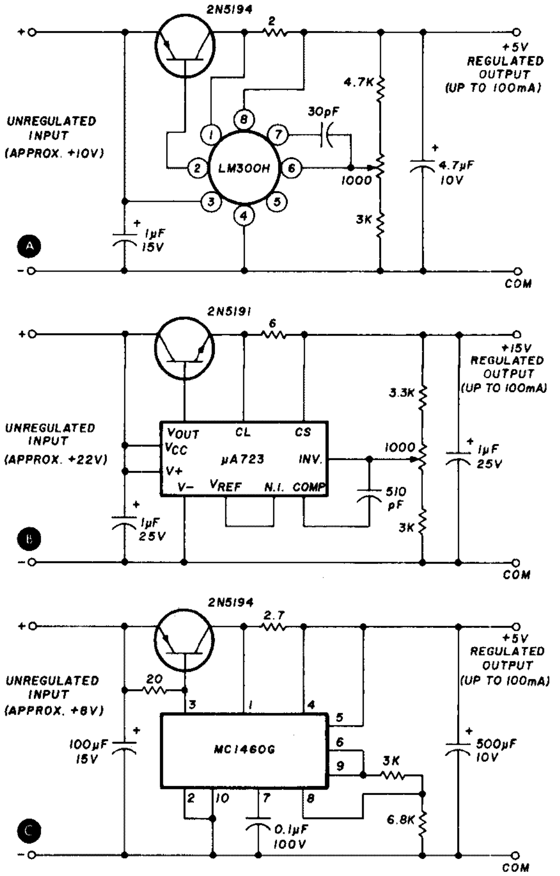


fig. 8. Series regulators using ICs especially designed for the purpose. All use external power transistors as part of the regulator circuit. The regulator in A uses the National LM300H or its improved version, the LM305H. In B the Fairchild μA723 is used. C shows another version using a Motorola 1460G, which has since been replaced by the MC1469. The National and Fairchild devices have become widely second-sourced and appear under a variety of brand names.

tor, because the increased ripple can be seen immediately. Bridging a new (correct value) electrolytic across the suspected open capacitor is a simple way to check this problem. What causes most electrolytic open failures is excessive ac. This current causes heat and subsequent drying of the capacitor electrolyte. If you have a supply in which electrolytes consistently dry up, probably no thought had been given to limiting ac current in the capacitor in the original design. This was a rare problem when we had only tubes and selenium rectifiers because of their high equivalent series forward resistance. But the modern silicon rectifier diode has extremely low equivalent forward resistance and immense peak current capability. A typical 1N4001 rectifier, rated at 1 amp, has a peak current rating of 30 amps!

With essentially no limit to the peak current our modern rectifiers will handle, the only limitations on ac in the filter capacitor, in a capacitor-input rectifier-filter system, are the leakage inductance and secondary resistance of the power transformer (plus the equivalent series resistance of the electrolytic itself). By using a clamp-on ac ammeter on the filter-capacitor lead you can easily see the magnitude of this current. The solution is either to use a capacitor of higher current rating (expensive and relatively bulky) or to add a small resistor, as in fig. 11. One-ohm, half-watt resistors are common and inexpensive for this purpose. For dc to 500 mA, this protective resistor will decrease the output less than 0.5 volt, which will usually be tolerated by the circuitry (regulator or load) that follows the rectifier filter. For larger-current supplies, a smaller resistance made from nichrome or constantin wire should be used. Note that the *exact* resistance and its stability are unimportant here as long as the resistance limits the ac current to a value the filter capacitor can accept according to its ratings.

magnetic components

The only other components in the power supply are the power transformer and perhaps a filter choke. These components usually fail from long-time overload or just plain old age and will have been "cooked." Obvious evidence of their demise includes strong odor, charred paper insulation, darkened insulation on formvar winding wire, and leaking insulation oil from sealed units.

Ohmmeter checks for winding continuity and leakage paths to the frame are called for here. Also ac voltage measurements across the power transformer secondary are appropriate. If all seems well in the rectifier-filter section, then the supply may be loaded using appropriate power resistors or light bulbs to determine if failure occurs only under load. The transformer alone may also be checked under resistive loads in the same way.

regulators

If you have problems in the regulator section, fixing becomes more varied and interesting. It can be very

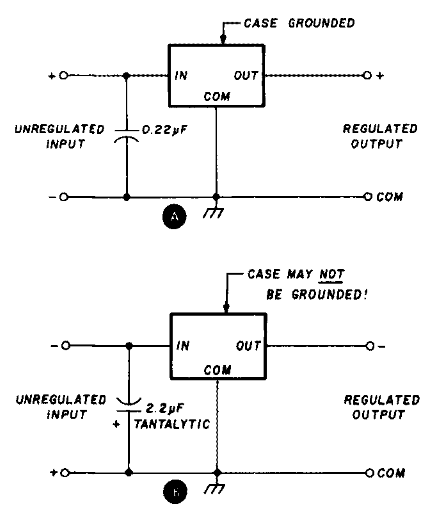


fig. 9. Using a three-terminal device as a positive regulator, A, and as a negative regulator, B, the power transistor is included within the chip. The IC for A can be a National LM340 or Fairchild μA7800; for B is can be a National LM320 or Fairchild μA7900. The capacitor between input and common is used for stability.

simple if the regulator is one of the three-terminal types, in which case the regulator is simply replaced. Regulator failures in my experience fall in this order of probability: series-pass power transistor, regulator IC, small transistors, and small electrolytics. Since the regulator is a feedback system, voltage measurements are often misleading. Unlike open-loop circuits, feedback systems usually completely tie themselves into knots when any part of them goes wrong.

The easiest way to fix a regulator is to check (or replace) the most likely parts (as above), one at a time, until success is achieved. Two voltage measurements can be helpful – sometimes: measurement across the current-sense resistor and across the reference. In fig. **7D** this would be across the 6-ohm resistor for current-limiting; any voltage here in excess of 0.6 volts indicates that the regulator is current limiting. Also in fig. **7D**, a check across the zener (1N4611) will show if the regulator has any reference voltage to which it can compare the output voltage.

Reference-zener failure doesn't occur often in good commercial designs, but a number of companies use inexpensive plastic transistors as reference diodes. The reverse breakdown voltage of the emitter-base junction of a silicon transistor is used for a zener in the 4-10 volt range. These "reference diodes" are inexpensive for the manufacturer but nasty to replace, because the repairman has no idea of their exact breakdown voltage before failure. My practice is to use a good zener with a sharp knee to replace these so-called references. The 1N4099 at 6.8 volts is my favorite at least as a first cut.

mysterious part labeling

When fixing commercially made power supplies you may encounter that annoying facet of protectivism called "in-house-numbering." For example, Motorola makes an MC1466L for Lambda to use in their modular power supplies *but gives it a* Lambda special number. There is *no way* you can find out from the IC manufacturer what the part is – short of industrial espionage. All

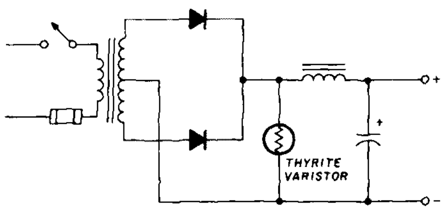


fig. 10. Choke-input circuit with a thyrite varistor added to protect rectifiers against overvoltage when the choke field collapses.

IC manufacturers do this for large O.E.M. customers. However, the IC manufacturer usually puts his trademark on the IC or transistor so that at least you have a clue.

There is almost zero probability that an in-house semiconductor is a specially-made device because of the cost factor; almost always it will be a standard part but re-marked. The circuit and pinning of the device to be replaced should be compared with the information in the semiconductor manufacturer's data book, then usually you can infer what the part is. *In modular* power supplies the Fairchild μA723 and National LM305 are by far the most common regulator ICs.

The main producers of regulator ICs are Fairchild, National, Motorola, Silicon General, Raytheon, Tele-

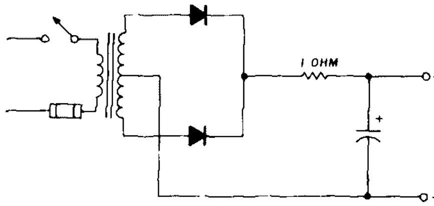


fig. 11. Capacitor-input filter with a 1-ohm resistor added to limit electrolytic-capacitor current.

dyne, Signetics, and RCA (that is, all the big linear houses). These companies' trademarks are easily recognized and their data books are commonly available. Occasionally you may be able to obtain an in-house number equivalent by calling the O.E.M. or his field engineering office. When replacing the regulator IC I always *carefully* unsolder the package using a desoldering tool (vacuum type) then *install an IC socket.* This allows a new IC to be tried without further weakening the circuit-board traces and pads.

heat sinks

One last point on power supplies is the subject of heat sinks. Very often commercial designs at the low-price end have inadequate heat sinks for power semiconductors. Occasionally you'll even find power semiconductors "heat sink" mounted to a steel chassis or bracket. This, of course, saves the O.E.M. money but provides *poor* heat conductivity. A simple retrofit of an aluminum or copper plate can often prevent persistent semiconductor failures due to overheating. Another pitfall is the *application* of modular power supplies, especially the molded-in-epoxy block types. These modules are frequently designed to be operated with good thermal contact to a large aluminum chassis or other good heat dissipator. Floating them on a fiberglass circuit board in a stagnant air location is a sure way for them to fail.

conclusion

We have touched on a number of types of power supplies and the ways in which they can be separated into sections, the section at fault isolated, and finally that section repaired. There are many other regulator forms: shunt regulators, switching regulators, and even regulators that rely on ferro-resonance in transformers and other clever magnetic tricks. These other regulators are not so common that space can be devoted to them here; but they, too, can be fixed using the same general principles: separate, isolate, repair.

ham radio

binaural synthesizer-filter with tone-tag modulation

A narrowband tone-marking system that quickly locates binaural crossover and provides a characteristic tonal quality to CW signals

Back in the thirties, before the explosion of electronic-component availability and refinements, amateur radio was more relaxed. You could tune across the phone bands and hear anything from a rock-crusher with broadcast-band quality eating up 20 "kilocycles" of band-width to a self-excited oscillator modulated by a loop-coupled telephone mike. CW was more often effected by a straight key than a bug. And since voltage regulators were not yet generally available, many CW signals chirped, thumped, clicked or yooped – some were pure raspberry (filter capacitors were expensive).

By Don E. Hildreth, W6NRW, Post Office Box 3, Sunnyvale, California 94088

Added to this charm was brass-pounder rhythms such as the "Banana Boat Roll" and "Lake Erie" swing and many, many more – most with generous amounts of syncopation thrown in. Individuality there was.

Today the phone bands are sideband-tight and log compressed. Almost all CW signals are chirpless dc, with or without crystal control. And, with a few exceptions,[1] electronic keyers coldly subtract character from keying. Clearly the effect of signal quality in days gone by was not all bad by any means. You could often copy a DX station precisely because he had the sound of a buzz saw

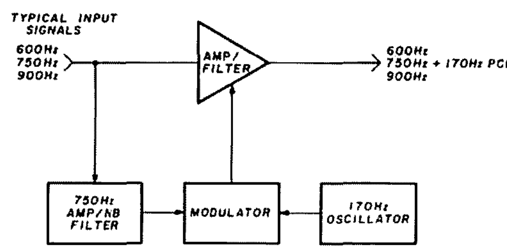


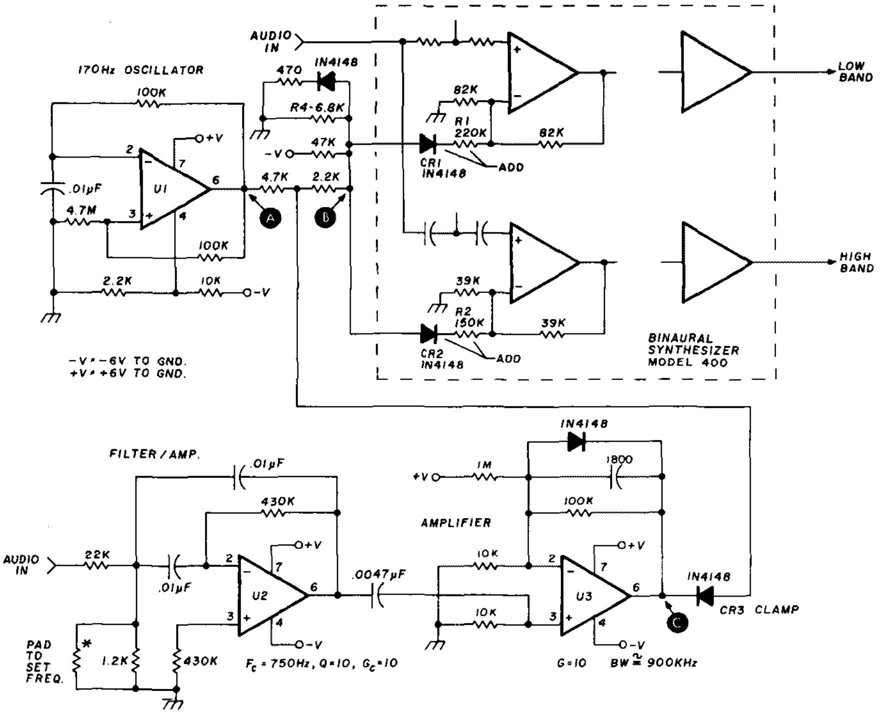
fig. 1. Block diagram of the basic Tone-Tag modulation system for enhancing reception of CW signals.

or maybe a peanut-whistle chirp. He was different, and it helped. Although some purists frowned on those raunchy sounds, they were less piercing to the ear (probably less damaging physiologically) and much more pleasant to copy.

system description

Because of technological progress much of the old

character of CW signals is gone forever; but perhaps today's technology can restore some of it. Of course you could make all signals sound like a buzz saw simply by cutting some of the power supply filter out of your receiver. You'd lose the piercing CW notes, but all signals would still sound alike. Tone-Tag™ is my answer to resistors R1 and R2 (fig. 2) between ground and open at a selected rate not only creates amplitude modulation, but it does so at the binaural crossover frequency. Every effort was made to obtain a modulated tone without superimposing startup delays or transients on the basic desired signal.



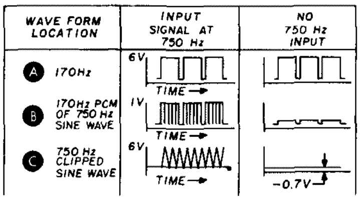


fig. 2. Circuit details of the basic Tone-Tag modulation scheme. Reducing resistance between an operational amplifier minus port to ground not only increased gain, but the gain is increased most near the filter cutoff frequency. The binaural synthesizer (dashed box) appeared in reference 2.

part of the problem. I use it to accomplish two major objectives:

1. Tone modulate a selected signal without changing adjacent signals.

2. Provide a tuning aid to determine quickly and precisely the location of the 750-Hz crossover point in my binaural synthesizer for CW.[2]

You'll note that the first objective can be accomplished whether or not you use a binaural synthesizer. You can apply the Tone-Tag principle to any system. The general idea is shown in **fig. 1**. **Fig. 2** shows circuit details. Although there are many ways to effect modulation, this method takes advantage of a basic characteristic of the VCVS active filter class. In this case, reducing resistance from a minus op-amp port to ground not only increases gain, but the gain is increased most around the filter cutoff frequency. Therefore, switching

circuit

A tone-modulating oscillator, U1, runs at approximately 170 Hz (this frequency is between the 4th and 5th subharmonics of the 750-Hz listening frequency to avoid harmonic beat notes). Under no-signal conditions, the modulating oscillator's assymetrical square wave is inhibited by clamp CR3. The clamp is held in conduction by a quiescent -0.7 volt output from op amp U3. Divider R3, R4 inhibits modulator diodes CR1, CR2 when strong signals pass through the system above and below 750 Hz. When a nominal 750-Hz signal appears, it is amplified and filtered by U2, a narrow-band filter, then fed to amplifier U3. Positive half cycles appearing at U3 output periodically open the clamp CR3, which allows samples of the 170-Hz oscillator to do its work on the modulation diodes. In this way, only those signals in the filter passband near crossover are modulated — a method that is as effective as it is simple. **Fig. 3** shows a

complete system including a 300-Hz input filter with adjustable skirts.

The 741 IC op amp or its equivalent will work in all circuits shown, but a 308 or 301A with a 3-pF frequency-compensation capacitor between pins 1 and 8 will provide more margin for the U2 circuit in fig. 2. Carbon-film resistors are rated better than carbon composition for the active filters, but 5%, ¼-watt carbon resistors are usually sufficient. It's better to use capacitors with a tolerance of 10% or better, unless you wish to spend some time with a bridge. I've had good results with 50- or 100-volt 10% mylar film capacitors (James Electronics and others).

function on input audio signals as low as 10 mV rms; therefore, considerable dynamic-range flexibility is provided. Hopefully your receiver, in addition to its other good features, has a well-designed i-f amplifier and detector (mixer for CW); for although an audio processor can help a marginal i-f system, it will do much better with one that provides a good single-signal response and one that does not generate copious quantities of spurs. An excellent bandwidth to work with this system is around 1 kHz with a 6-60 dB shape factor of 2.

operating procedures

With your receiver bfo set around 750 Hz above or

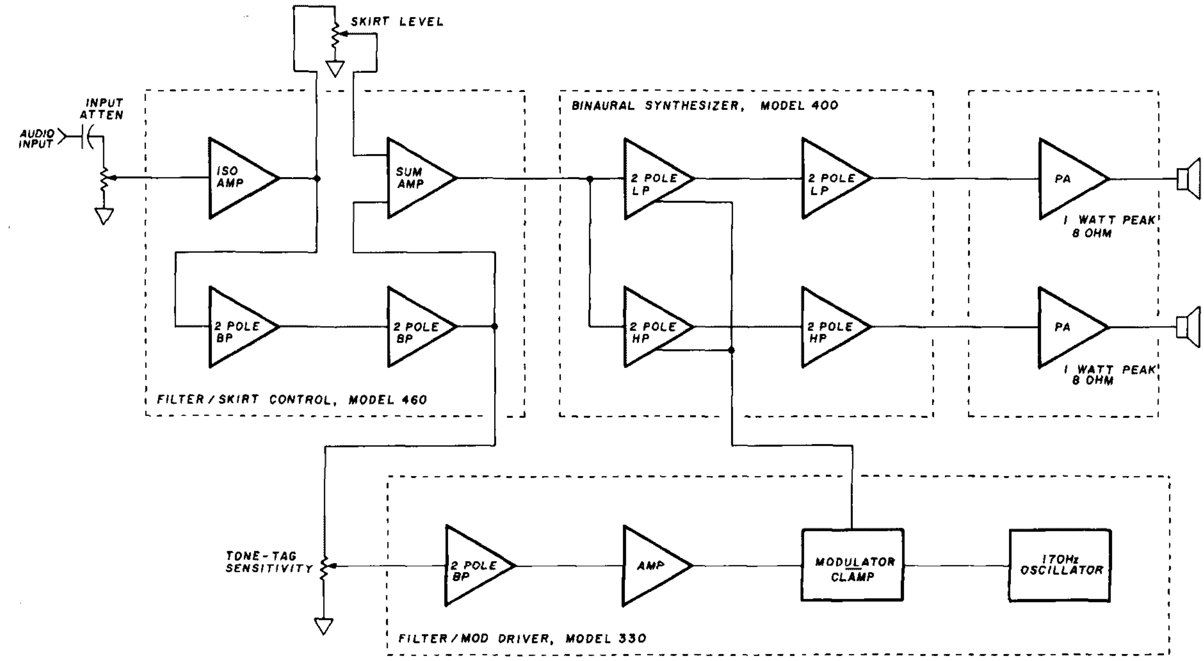


fig. 3. Complete Model 1100 system including Tone-Tag, binaural synthesizer, and 4-pole, 300-Hz input filter.

input requirements

Since distortion in the driving receiver and early stages of this system can reduce effective binaural action, very linear operation is desirable. This can be assured by keeping signals out of the final low- and high-pass active filter stages to a level of *not more* than 5 or 6 volts p-p. This level will also avoid interaction between signal levels and the modulation function.

As a distortion-effect example, if a receiver produces a condition where the second or third harmonics of a hypothetical 500-Hz signal are down only 10 or 20 dB, then the effect of the excellent binaural separation built into the system is largely compromised. A similar case is true for intermodulation distortion, and we have enough of these problems with our nonlinear ears. When you consider that the maximum physiological binaural potential tends to be only about 7 dB,[3] it's best to stay very linear (operate well below saturation) and don't listen with more than comfortable volume.

Available gain in the Tone-Tag system will allow it to

below its i-f passband center and with binaural skirts *up,* simply tune a desired CW station to binaural center. At this point, coincident with equal energy in your right and left speakers or phones, the received signal will pick up a gentle tone modulation. At the same time, signals to the right and left (higher or lower beat note) will remain pure dc. Under heavy interference you may turn the skirt-control to drop the wideband skirts, leaving the tone-modulated signal unchanged. Usually, however, it seems better to leave the skirts — or floor — reasonably well up and enjoy three modes of selectivity: binaural, Tone-Tag, and the "ear-brain" filter.[4] Frequency response with typical skirt control positions is shown in **fig. 4.**

What about using this system for receiving phone signals? To me they sound better than with a monaural audio system. The stations you hear seem to occupy a wide band of space in front of you; but of course, Tone-Tag just doesn't come into the picture. Under interference or high noise, you can drop the flat skirts 10 dB

or so to improve copy. With music, leave the skirts up and enjoy simulated stereo.

conclusions

With Tone-Tag, one of the desirable factors of old, lost through technological improvement, has been returned — at *your* choice of application. Now you can apply tone modulation *and* binaural action to any CW signal you select. This provides a much wider field than before. And while it is true you won't have much luck binaurally separating or modulating just one of several signals within a few Hz of each other, these cases are not too frequent. On balance, we are ahead.

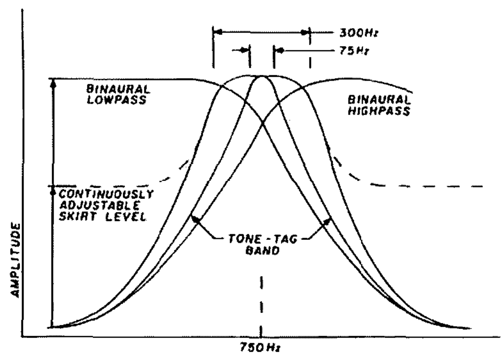


fig. 4. Frequency response of the binaural bandpass filter and Tone-Tag modulation system. The 75-Hz Tone-Tag filter is cascaded with the 300-Hz input filter, which provides skirts equivalent to those from a 6-pole filter in addition to enhancement from the binaural filters.

A narrowband filter is used as a part of this system; but in this case it's not in the direct signal path so it doesn't ring and ping with noise and signal alike. This system helps the selective process without interfering with our fantastic ear-brain filter.

Considerable experimentation was applied in choosing a tone-modulated frequency. Brainwave rates were even tried. "Alpha" sounds fine as long as keying rates are slow; but for all-around use, the frequency shown seemed best to me.

acknowledgement

I wish to thank Dick Turrin, W2IMU, for providing a copy of reference 4 and for some very interesting references indicating potential *s/n* improvements when comparing monaural with binaural listening.

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ham radio

calculating line-of-sight distance to the horizon

Users of the vhf/uhf spectrum often want to know the line-of-sight distance from the top of some particular object to the horizon. The distance is given by the following formula:

$$D = \sqrt{2Rh + h^2} \qquad (1)$$

where

D = line-of-sight distance from the observing point to the horizon

R = radius of the earth to the average terrain surrounding the observing point

h = height of the observing point above the average terrain

You may use any units of measure you like as long as you are consistent throughout the equation.

Eq. 1 does not take into consideration variations in terrain between the observer and the horizon, which of course can alter the actual distance that he will see, nor does it consider the characteristics of propagation (refraction, multipath), which often affect transmission range.

The value of R is not critical, however, regardless of latitude or altitude above sea level. The following approximations, though based on mid latitudes, will give excellent results regardless of where you live.

$$R \approx 20{,}890{,}400 \text{ feet}$$

or

$$R \approx 6{,}367{,}400 \text{ meters}$$

The really important factor is the height of the observing point above the average terrain, and this figure should be as accurate as possible.*

While the distance formula is quite simple, it is not always easy to remember the radius of the earth. For those who use the English system of measurements, the following is an excellent approximation:

$$D\,(miles) = \sqrt{1.5h}\,(feet) \qquad (2)$$

*Excellent topographical charts are avilable that show contour lines depicting elevation in feet above mean sea level for just about any community in the country. These charts are avilable at nominal cost from the U.S. Geological Survey, Denver, Colorado 80225, or Washington, D.C. 20242. A folder describing topographical maps and symbols is available on request from the USGS. **Editor**

By William D. Johnston, WB5CBC, 1808 Pomona Drive, Las Cruces, New Mexico 88001

Note that the height is in *feet* and the answer is in *miles*. This approximation is so close that the distance error does not exceed a tenth of a mile until you rise to well over 60,000 feet in altitude. (This also makes it useful in finding the distance to the horizon from an airplane). You may find it interesting to prove to yourself just why this approximation is valid.

Conversely, if you want to know how high you must be to see a certain distance, **Eq. 2** is rewritten as:

$$h \approx \frac{2D^2}{3} \qquad (3)$$

Once again, remember that distance is in miles and height is in feet.

If you use the metric system of measurements, the following is an excellent approximation for line-of-sight distance to the horizon in *kilometers* when the height is in *meters:*

$$D\,(kilometers) = \sqrt{12.7h}\;(meters) \qquad (4)$$

To see how high you must be to see a certain distance, this may be rewritten as

$$h\,(meters) = \frac{D^2}{12.7}\;(kilometers) \qquad (5)$$

Table 1 will give you an idea of what the formulas reveal for both metric and English measurements. You may be surprised that, even at fairly high altitudes, the horizon really is not very far away. As a friend of mine said recently, "Not only is the earth round but it is very, very round."

table 1. Distance to the horizon as a function of observing height, English and metric measurement systems.

observing height, h (ft)	distance, D (mi)	observing height, h (m)	distance D (km)
1000	39	300	62
2000	55	600	87
3000	67	900	107
4000	77	1200	124
5000	87	1500	138
6000	95	1800	151
7000	102	2100	164
8000	110	2400	175
9000	116	2700	185
10000	122	3000	195
11000	128	3300	205
12000	134	3600	214
13000	140	3900	223
14000	145	4200	231
15000	150	4500	239

ham radio

microcomputer interfacing: a software UART

This month we return to the subject of the substitution of software for hardware, i.e., the substitution of machine-level routines and subroutines for specific digital hardware devices that store, manipulate, transmit, or receive digital information. The hardware device we'll discuss is the universal asynchronous receiver/transmitter, or UART – a 40-pin integrated-circuit chip that contains an independent 8-bit asynchronous receiver and an independent 8-bit asynchronous transmitter. Data rates range from dc to 60,000 bits per second. The receiver and transmitter sections of the chip can be programmed for 5, 6, 7, or 8 data bits; 1 or 2 stop bits; even or odd parity; and parity or no parity. The chip also contains a variety of flags. For further details, we refer you to manufacturer's literature or to references 1 through 3.

UART interface

An interface circuit for a simplified software UART is shown in fig. 1. Because of the nature of the specific application, which we'll discuss at the end of this column, there was no need for special flag bits or error checking. As a consequence, the interface circuit consists of a single three-state input buffer gate (SN74126), a single output-data latch (SN7474), two input device-select pulses, and one output device-select pulse. With appropriate modifications of the device-select pulses, this circuit can be used with almost any microprocessor chip. In our case, a 8080A-based microcomputer operating at 750 kHz was used. This system generates and detects, in combination with operating software, asynchronous serial ASCII-coded, 5-volt, TTL data. For teleprinter operation, additional hardware is required to convert the 5-volt logic levels to 20 mA current loop operation.[4]

By Paul E. Field, David G. Larsen, WB4HYJ, Peter R. Rony, and Jonathan A. Titus

Dr. Field is guest author of this month's column. Dr. Field and Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Titus is President of Tychon, Inc., Blacksburg, Virginia.

The transmit subroutine, shown in table 1, for the software UART occupies 20-25 successive program steps in memory once the appropriate PUSH, POP, and RET instructions have been included. Also required is a 9.09 millisecond time-delay subroutine, which corresponds to an asynchronous serial ASCII data transmission rate of 110 Baud, i.e., teleprinter speed. The program in table 1 is described as follows:

Register L is used as the bit counter for the 11-bit ASCII word, and is set initially to octal 013. The seven data bits plus the parity bit, which is bit 8, are assumed to be present in the accumulator. At LO = 146, the accumulator is ORed to itself to clear the carry bit, which is shown on the far left in fig. 2. In fig. 2 the least-significant data bit is bit 1. At address LO = 147, a RAL instruction is performed to rotate the start bit to bit position DO in the accumulator. Fig. 3 should pro-

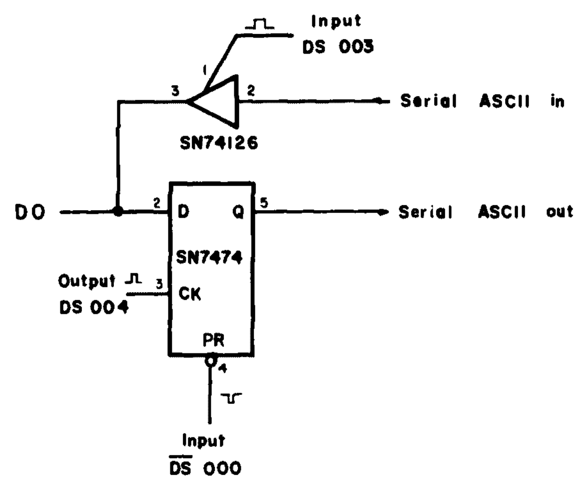

fig. 1. Interface circuit between an 8080A-based microcomputer and a TTL asynchronous serial ASCII input-output device.

vide assistance in understanding the four different rotate instructions in the 8080A microprocessor instruction set.[5]

At LO = 150, the start bit is output to the SN7474 data latch, which is shown in **fig. 1**. The program then goes into a 9.09 ms time delay subroutine, after which bit 1 is rotated into the DO accumulator position and the carry bit is set to logic 1. Bit **1** is output to the SN7474 latch, the ASCII word bit counter in register L is decremented, and program control is returned to the time delay subroutine, which is called at LO = 152. The loop from LO =152 to LO = 164 is executed a total of eleven times, after which register L becomes zero and the JNZ instruction at LO = 162 is ignored. A software UART transmit subroutine possesses a flexibility equivalent to the original 40-pin UART chip. With appropriate modifications to the program or the original accumulator data, you can transmit 5, 6, 7, or 8 data bits; 1 or 2 stop bits; even or odd parity; and parity or no parity.

table 1. Microcomputer subroutine that demonstrates the asynchronous serial transmission of an eleven-bit ASCII word at a teletype speed of 110 Baud.

LO memory address	instruction byte	mnemonic	description
•	•	•	
•	•	•	
•	Accumulator contains 8-bit ASCII word. Bit 8 is the parity bit, which can be set for even or odd parity, or no parity.		
144	056	MVI L	Set ASCII word bit counter to 013
145	013	013	
146	267	ORA A	Set carry bit to logic 0
147	027	RAL	Rotate carry bit to DO in accumulator
150	323	OUT	Output carry bit to SN7474 latch
151	004	004	
152	315	CALL	Call 9.09 ms time delay subroutine
153	<B2>	<B2>	LO address byte of time delay subroutine
154	<B3>	<B3>	HI address byte of time delay subroutine
155	037	RAR	Rotate bit in ASCII word to DO in accumulator
156	067	STC	Set carry bit to logic 1
157	323	OUT	Output bit to SN7474 latch
160	004	004	
161	055	DCR L	Decrement bit counter by 1
162	302	JNZ	If bit counter has value of zero, ignore this instruction. If all of the bits in the 11-bit ASCII word have not yet been transmitted, jump to address LO = 152 above.
163	152	152	LO address byte
164	<B3>	<B3>	HI address byte
•			
•			
•	At this point, the 8-bit ASCII word contained in the accumulator has been transmitted. Two stop bits have been added at the end of the eight bits and a single start bit, at logic 0, has been added at the beginning of the eight bits.		
•			

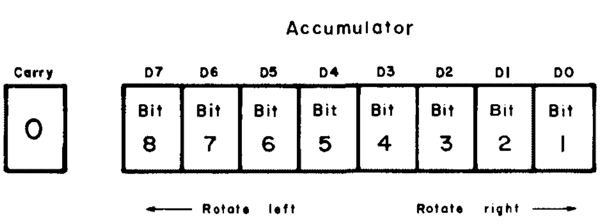

fig. 2. Schematic of the accumulator contents after the ORA A instruction in the software UART transmit subroutine in table 1. The carry bit is the start bit, which is at logic 0.

The time-delay subroutine can be modified so that you can transmit at data rates from 60 to 9600 Baud for a 750-kHz clock rate and higher for 2-MHz and 4-MHz 8080A clock rates.

The conversion from one data transmission rate to another is easily accomplished with the aid of appropriate software time delay subroutines, which replace RC time-constant circuits. An additional advantage that accrues from the use of software is the potential to perform code conversions. For example, 5-level Baudot KSR machines are in widespread use and can still be obtained for under $50. It is not too difficult to develop software that converts ASCII to Baudot and thus produce an inexpensive hard-copy terminal for the laboratory scientist or engineer, amateur, or computer buff.

The software UART receive subroutine requires 50 instructions and will not be repeated here.* The basic programming concepts associated with the receive subroutine are illustrated in fig. 4, which represents an eleven-bit asynchronous series ASCII word that is being detected by the 8080A-based microcomputer with the aid of the SN74126 three-state buffer gate shown in fig. 1. The program repeatedly tests the "serial ASCII in" line in fig. 1 for a logic 0 state. Once a logic 0 state, which corresponds to a start bit, is detected, the program goes into a 4.54-ms wait loop. Upon leaving the wait loop, the program again inputs the logic 0 into bit

*Copies of the transmit and receive subroutines and a description of a "smart" remote-data entry station are available from Prof. Paul Field, Department of Chemistry, Virginia Polytechnic Institute and State University, Blacksburg, Virginia 20461.

position DO in the accumulator, thus testing the validity of the start bit. The start bit is rotated to the carry bit, and the program then enters a 9.09-ms wait loop, after which it inputs bit 1 into position DO in the accumulator. Register H is used as the SAVE register, which stores the growing ASCII data word. The SAVE register is rotated one position, and the 9.09-ms wait loop is again entered, after which bit 2, which is a logic 0 in fig. 4, is input into bit position DO in the accumulator. The input of successive data and parity bits continues until the entire 8-bit data word is entered into the SAVE register. The two stop bits are also detected. With appropriate modifications, the program can detect parity or framing errors or an overrun condition. A data ready flag signal can also be generated from software with the aid of a second SN7474 latch.

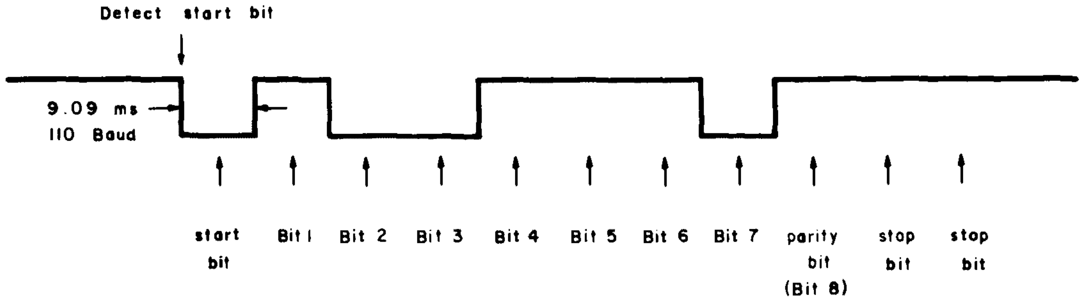

fig. 4. Schematic of an eleven-bit asynchronous serial ASCII word being received by the 8080A-based microcomputer. Each bit has a time duration of 9.09 ms. The logic state of each bit is detected at the middle of the bit.

The software UART routines described above by Dr. Paul Field of VPI & SU were in a "smart" remote data entry station that was tied through a 20-mA current loop to a PDP 8/L minicomputer in a physical chemistry laboratory. The data entry station intercepted the 20-mA teletype current loop tied to the minicomputer. The remote data entry station permitted students to load data into memory and then transmit it as a block to the minicomputer, which analyzed the data and provided a printout. With the aid of the 20-mA current loop operated in the full duplex mode, ten or more remote data entry stations could be tied to the minicomputer.

This column provides a good demonstration of the software-hardware tradeoffs that can be accomplished using microcomputers. Similar, and perhaps more comprehensive, routines have already been written for all the popular microprocessor chips, such as the 16-bit PACE or the 8-bit 6800. The faster and less expensive microcomputers become, the more likely that all moderate-speed digital functions will be executed through software. The theme of software replacing hardware is an important one, and we'll return to it many times in the future.

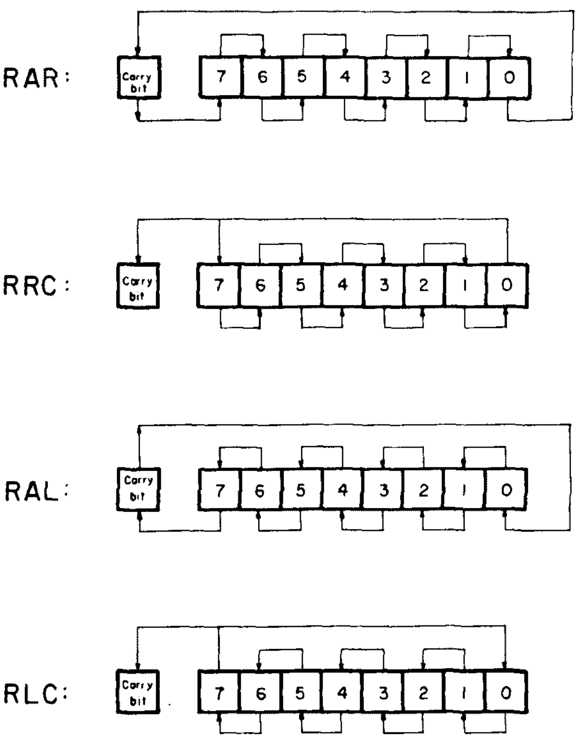

fig. 3. The four rotate instructions in the 8080A microprocessor instruction set.

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1. *Bugbook IIA, Interfacing and Scientific Data Communication Experiments Using the Universal Asynchronous Receiver/Transmitter (UART) and 20 mA Current Loops,* E & L Instruments, Inc., Derby, Connecticut, 1975. (Available from Ham Radio Books, Greenville, New Hampshire 03048. Price $4.95; Order no. BB-2A).

2. D.G. Larsen and P.R. Rony, "Computer Interfacing: INWAS, Interfacing with Asynchronous Serial," *American Laboratory*, 6 (9), 83, 1975.

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4. P.R. Rony, D.G. Larsen, and J.A. Titus, "Computer Interfacing: Half Duplex and Full Duplex Current Loops," *American Laboratory,* 7 (6), 78, 1975.

5. *Bugbook III, Microcomputer Interfacing Experiments Using the Mark 80 Microcomputer, and 8080 System,* E & L Instruments, Inc., Derby, Connecticut, 1975. (Available from Ham Radio Books, Greenville, New Hampshire 03048. Price $14.95; Order no. BB-3.)

ham radio

the ham notebook

ten-minute timer

The SN74121 works well as an astable multivibrator although it was designed for monostable operation. In the circuit of fig. 1, U1 generates pulses at 4-second intervals. U2 and U3 divide the pulse train by 144 which results in a period of 576 seconds (9 minutes, 36 seconds). U4 is then turned on and produces a positive output pulse of 20 seconds duration. Transistor Q1 actuates a keyer or other signalling device such as a sidetone oscillator, lamp or LED.

The LED connected to pin 1 of U1 flashes a few milliseconds at the beginning of each 4-second period and makes it easy to adjust the time interval in comparison with WWV. The timing adjustment is made with trimpot R1. The timing range may be extended by changing R1 to 5 kilohms and R2 to 45k.

It is possible to reset the counter (see dashed circuitry), but this requires an additional switch. At my station, I identify at the beginning of a QSO, and again when the ten-minute alarm flashes — then I am in synchronization with the timer.

The timing tolerance over the full period is within about ±5 seconds. However, this requires good quality capacitors at C4 and C5 and a stable power supply.

Herbert Seeger, DJ9RP

audio mixer

I recently came up with a system which materially improved my ssb "talk power" and am offering it here to the amateur fraternity. I had been using an Astatic D-104 microphone with relatively little success in the DX pileups. Friends told me the audio sounded "thin" so I bought an Electrovoice EV-674 dynamic. Reports with the EV-674 indicated I sounded smooth but muffled, as if talking into a barrel. It occurred to me that the best features of each microphone could be combined with an audio mixer. Radio Shack sells a well-made, four channel transistorized mixer for $13.95.

I use two of the channels and mix the D104 with the EV-674 in the proper ratio. My ssb signal now has audio punch and quality. I also get through the pileups sooner. The "W6KNE Equalizer" should help provide good audio for everyone, particularly those operators whose natural voice does not come across strong and with "punch."

Gary Legel, W6KNE

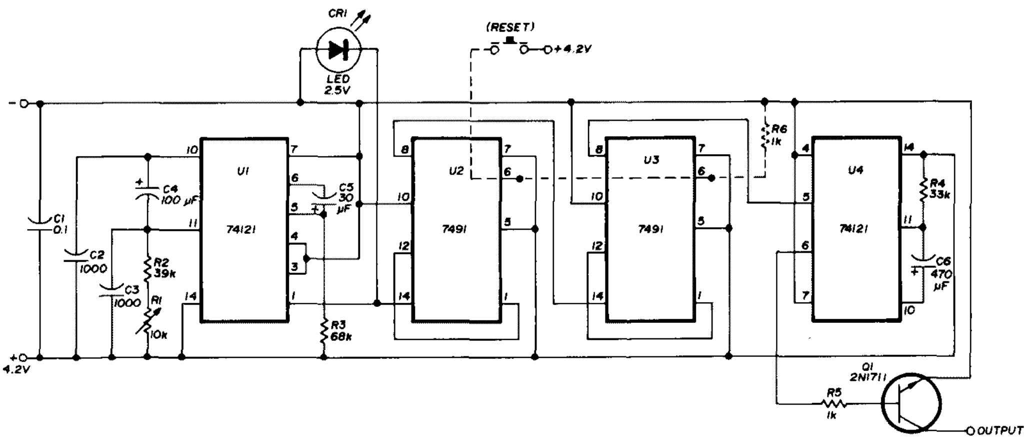
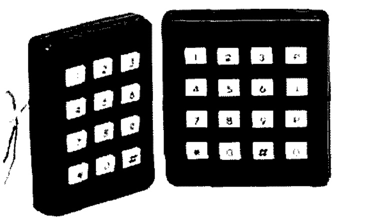


fig. 1. Ten-minute timer uses 74121 (U1) as an astable multivibrator, generating pulses a 4-second intervals. U2 and U3 divide the pulse train by 144 which results in a period of 576 seconds (9 minutes, 36 seconds).

NEW products

new touch-tone encoders for variety of applications



Looking for a *Touch-Tone* encoder? Need 12 or 16 buttons for mobile, hand-held or desk-mounting applications? If so, Pipo Communications may have exactly what you've been looking for in their recently announced PP-1 (12-button) and PP-2 (16-button) *Touch-Tone* encoders.

The PP-1 measures just 3/8x2x2-11/16 inches (9.5x50.8x68mm) and the PP-2 measures 3/8x2-3/8x2-11/16 inches (9.5x60.3x68mm). Both pads are self-contained and feature positive touch control; that is, the keys depress. There are no "potted" parts, so servicing is easy, and the units are immune to rf interference. They possess low distortion at high output levels while drawing current only when the buttons are depressed. The output level is easily adjustable from the front of either pad.

Each unit is supplied with instructions for hook-up and use, schematic diagram, drilling template and all necessary mounting hardware.

These *Touch-Tone* encoders are capable of operating at any temperature between zero and 140 degrees F., at any voltage between 4.5 and 16 volts dc, and draw between 7.5 and 20 mA in use.

The PP-1 is priced at $55.00, and the PP-2 at $58.00. (California residents please add 6% sales tax.)

For additional information, write Pipo Communications, Box 3435, Hollywood, California 90028; telephone (213) 852-1515, or use *check-off* on page 126.

new catalog from Heathkit

Heathkit has just announced its colorful new fall 1976 catalog, a free 96-page release describing nearly 400 electronic kits for virtually every do-it-yourself interest. Heathkit product categories include amateur radio, hi-fi components, color TV, test instruments, digital clocks and weather instruments, radio control equipment, marine, aircraft and auto accessories, and many more.

Some of the interesting new products described in the fall catalog are: a new electronic TV game for Heath solid-state TV owners, a combination digital clock/indoor-outdoor thermometer, a portable power megaphone/yelp alert, professional-quality harmonic and IM distortion analyzers, a shirt-pocket-size color alignment generator, and a touch-control light switch.

Heath Company is the world's largest manufacturer of electronic products in kit form. Their step-by-step instruction manuals can be followed easily by anyone, and are world-famous for clarity, precision and accuracy.

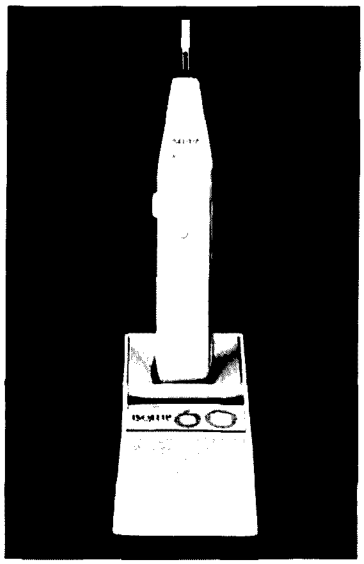
For your new, free, catalog write Heath Company, Department 350-04, Benton Harbor, Michigan 49022; telephone (616) 982-3417, or use *check-off* on page 126.

alphanumeric printer kit

Southwest Technical Products Corporation recently announced its new low-cost alpha-numeric printer kit. The PR-40 is a 5x7 dot matrix impact printer capable of printing the 64-character upper case ASCII set with 40 characters/line at a print rate of 75 lines/minute on standard 3-7/8 inches (9.84cm) wide rolls of adding machine paper. One complete line is printed at a time from an internal 40-character line buffer memory. Printing takes place either upon the receipt of a carriage return, or automatically, whenever the line buffer memory is filled.

The PR-40 printer is available in kit form only and includes the assembled print mechanism, chassis, circuit boards, components, 120/240-volt ac, 50/60 Hz power supply, assembly instructions, one ribbon and one roll of paper. The price is $250, postpaid, in the United States. Delivery is approximately 30 days. For more information write Southwest Technical Products Corporation, 219 West Rhapsody, San Antonio, Texas 78216; telephone (512) 344-0241, or use *check-off* on page 126.

rechargable cordless soldering iron



Wahl Clipper Corporation recently introduced its new *Iso-Tip 60* Cordless Soldering Iron. The low voltage, battery-operated, ground-free unit is the first that can be recharged from "dead" to "full" in one hour. With the *Iso-Tip 60*, the user can enjoy virtually uninterrupted service from a single unit, making cordless soldering practical for heavy-use applications.

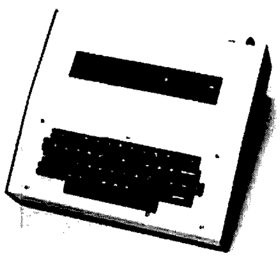
The *Iso-Tip 60* Cordless Soldering Iron has the capacity for up to 125 or more electronic joints on a single

charge. It can, however, be kept at a constant "full" charge simply by resting it in the recharge stand when not in use. The iron is equipped to accept a very fast charge when its battery is down and then switch to a trickle rate for maintenance purpose. An LED indicator shows when the unit is fully charged.

Standard kit no. 7800 from Wahl includes an *Iso-Tip 60* Cordless Soldering Iron, a recharging stand, a fine tip, a chisel tip and an instruction booklet. Any of the 16 Wahl snap-in soldering tips are usable with the new iron.

For further information contact Ross Advertising, Inc., 5901 N. Prospect Road, Peoria, Illinois 61614, or use *check-off* on page 126.

KSR-type terminals from Info-Tech



Info-Tech, Incorporated, a manufacturer of digital electronic systems for the amateur radio market and for light computer use, recently announced two new KSR-type terminals featuring low price and versatility. One terminal features 16 lines of 32 characters and the other, 16 lines of 64 characters; both with RS-232 compatibility.

The new terminals interface with all popular micro-computer kits and any video monitor.

Detailed technical information and prices may be obtained from Info-Tech, Incorporated, 20 Worthington Drive, St. Louis, Missouri 63043; telephone (314) 576-5489 or use *check-off* on page 126.

ultra low noise jfets

National Semiconductor Corporation announces a new series of ultra low noise junction field-effect transistors. In the past, special selection by the manufacturer was required to provide ultra low noise parts, a factor contributing to their relatively high price and limited availability.

The new series of National jfets is specifically produced for ultra low noise audio and video applications, including particle detectors, vidicon and I-R sensor preamplifiers, and audio and video tape amplifiers. The series includes three metal-can devices (NF5101, NF5102 and NF5103) and three TO-92 epoxy devices (PF5101, PF5102 and PF5103).

Key specifications for the new fets include a typical noise figure of 1.5 dB at 10 Hz, and a common-source transconductance of 4000 micromhos minimum with a drain current of 0.5 milliamperes. Another feature of the new

series is tight control of the gate-source cutoff voltage. The NF5101 and PF5101 have a gate-source cutoff voltage range of 0.5 to 1.0 volts; the NF5102 and PF5102 have a gate-source cutoff voltage range of 0.8 to 1.4 volt and the NF5103 and PF5103 have a gate-source cutoff voltage range of 1.3 to 2.5 volts.

Unit prices range from about $1.00 for the plastic version and $2.50 lor the metal package. For further information, including electrical and performance characteristics of these devices, write to National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95051 or use *check-off* on page 126.

Allied Electronics catalog

Off press and ready for mailing is Allied Electronic's *1977 Engineering Manual and Purchasing Guide.* The guide is filled with a wide selection of industrial-type electronic parts, components, supplies, and equipment which keep amateurs, engineers, technicians, and hobbyists requesting Allied's guide year after year. And, new in this year's guide is the introduction of metric measurements on many electronic parts and components.

The 1977 edition of the guide is 212 pages full of high-quality electronic parts and equipment from Allied and other leading manufacturers. Choose from a wide variety of new products, in addition to the traditional items which have set the standard for Allied's previous guides. You'll find wire, cable, solid-state devices, test equipment, resistors, trimmers and potentiometers, transformers, switches, timers, connectors, relays, tools, capacitors, new solar energy products, test equipment and even a microcomputer, plus many other electronic parts too numerous to list. Allied offers bulk pricing for quantity buyers, and six nationwide stocking locations assure prompt delivery of ordered merchandise.

There's no need to rely on a friend or business associate for access to Allied's 1977 guide. Write today for your own copy. Send $1.00, to help cover postage and handling, to Allied Electronics, Dept. 77, 401 East 8th Street, Fort Worth, Texas 76102.

More Details? CHECK—OFF Page 126

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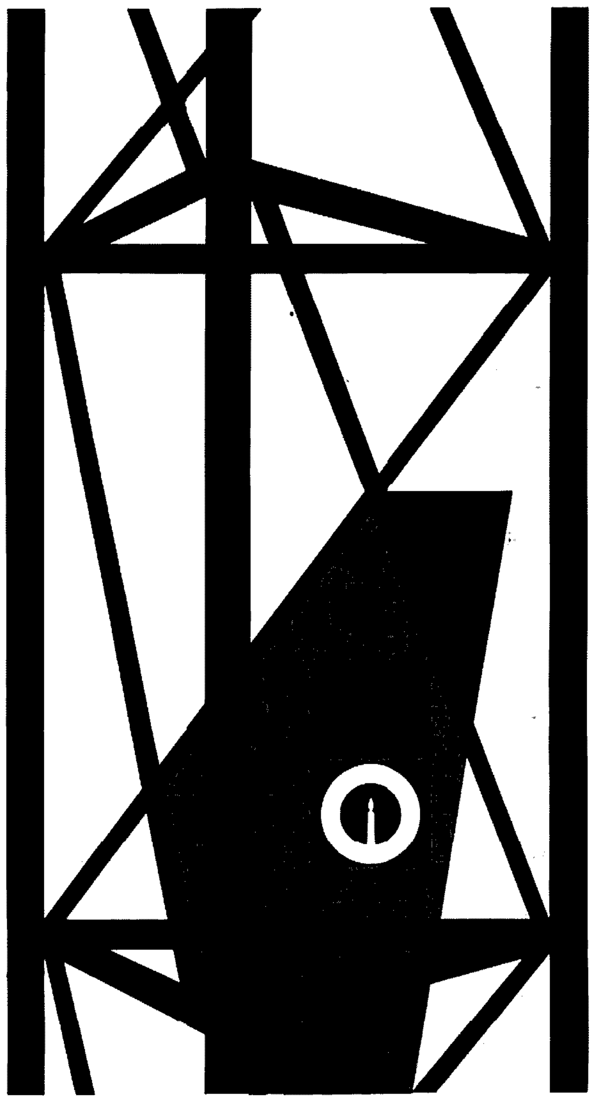
radio

DECEMBER 1976



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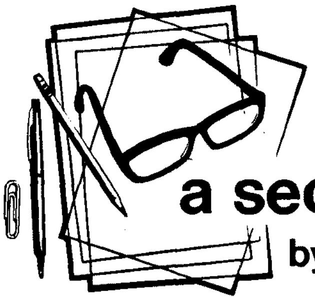


a second look

by Jim Fisk

In looking back over the advances in IC technology that have occurred during the past several years, it's hard to realize that integrated circuits actually celebrated their 18th birthday this past summer. It was during the summer and fall of 1958 that Jack S. Kilby of Texas Instruments built the first integrated circuit. Other semiconductor companies had been working on ways to miniaturize solid-state electronic circuits, but most of these techniques used miniature components of one kind or another; Kilby was the first to use semiconductor material for both the active (transistors) and passive elements (resistors and capacitors) to build a complete circuit on a single piece of germanium.

His first circuits, a phase-shift oscillator and multivibrator, demonstrated the feasibility of this approach. Since, at that time, germanium was well established as a semiconductor material, and silicon was not, Kilby used germanium. On top of the germanium wafer were the contacts of the diffused transistors, junction capacitors, and resistors. A gold-plated metal frame protruded from the lower surface of the substrate, and thermally-bonded gold wires were used for connections between those elements not linked by the wafer itself.

Kilby's first circuits were large and irregular – at least by today's standards – and were considerably different from the precision ICs that are presently on the market. The photo masks and resists necessary to IC manufacturing were yet to be developed, so the patterns were hand painted on the semiconductor chip with black wax. Needless to say, the end product was rather crude looking (and huge by today's standards), but it worked.

About the same time Kilby was working on his first integrated circuits, Fairchild Semiconductor developed the *Planar* process – an innovation that is generally conceded to be the foremost semiconductor discovery of the 1960s. This process made semiconductors more reliable and cheaper to produce, as well as accelerating IC progress and acceptance.

Since these early developments, the number of circuits per unit area has increased and prices have plummeted. In 1962, a typical IC flip-flop chip was about 0.1 inch (2.5mm) square; by 1970 a similar circuit was one-tenth that size, and today the same circuit is even smaller. At the same time circuit size (and unit cost) were decreasing, circuit speed increased from audio to 1 MHz or so, then to 5 MHz, 15 MHz, and on up the scale. It was only a few short years ago that a 30-MHz frequency counter was only a dream – today 600-MHz counters are commonplace, and even higher count frequencies are available if you're willing to pay a small premium for the capability.

It wasn't all that long ago that I reported on this page that, ". . . you can now buy a dual flip-flop for a couple of dollars or a complete decade counter for about seven." And these were RTL devices, with maximum counting speeds in the low MHz range. The same issue carried ads for 709 op amps at $3.98 and 711 dual comparators for $4.98. A low-current voltage-regulator IC, if you could find one, cost ten dollars or more. Scanning through the ads in the back of this issue, you can now buy a 30-MHz decade counter for about 45 cents, a 709 op amp for 29 cents, and a 711 dual comparator for 39 cents – about one-sixteenth of their 1968 prices. Considering that the consumer price index has increased nearly 70 per cent during the same period of time, ICs have to be among the best buys of all time.

Although I've said it before, it's worth saying again that, with the sophisticated, low cost ICs that are on the market, it's possible for amateurs to build exotic electronic equipment that only large laboratories with big budgets could afford a few years ago – and some they couldn't afford at any price because the technology just wasn't available!

Jim Fisk, W1DTY
editor-in-chief

presstop

"N" PREFIXED TWO-LETTER CALLS are now available to any Amateur eligible for a 1x2 callsign. The release of N prefixes applies only to 1x2 callsigns at this time, though presumedly the N prefix will also continue to be available when requested for special events stations.

"X" Suffix Callsigns, released in late October, increases total pool of 1x2 calls in each district by 676. Although the X suffix has traditionally indicated an experimental station in the United States, 1x2 X calls have not been issued since well before World War II so are not expected to cause any great confusion. Who's first for W1XX, Whiskey one Double Cross?

AMSAT'S ANNUAL MEETING drew about 75 people to Goddard Space Flight Center in October. JA1ANG brought the prototype 2-meter to 70-cm transponder from JAMSAT, and reported about 500 Japanese Amateurs are users of OSCARs 6 and 7. The AMSAT 2-10 meter transponder for OSCAR A-OD is now ready for test.

OSCAR 6's Birthday was October 15 and AMSAT's longest lived bird is over four years old and still going strong. As the launch anniversary ended OSCAR 6 began its 18,300 orbit, a lot of miles and many thousands of contacts to its credit.

OSCAR 7 Also celebrated a birthday; it was a healthy two-years old November 15.

N6V's Operation by the Jet Propulsion Laboratory's Amateur Radio Club was seen by millions of viewers on NBC's Today show. Featured during the outstanding segment was N6V's relaying of Mars photographs throughout the world via Amateur slow-scan TV.

The European Space Agency has announced approval of an Amateur Radio satellite launch on its experimental Ariane launch vehicle. A German-built OSCAR is scheduled to go aloft with Ariane's test flight B during December, 1979.

EXCELLENT RFI/EMI SURVEY article appeared in the September 20, 1976 issue of Electronic Design, should be "must" reading for any one seriously interested in the subject. Interference to as well as by communications equipment is included, and there's strong emphasis on FCC's increasing role in the problem area.

CB Is A Major cause of TVI/RFI complaints, of course, and in a recent discussion with FCC Chief Engineer Ray Spence he made the interesting observation that by far the majority of complaints of CB-caused interference was found to involve illegal power!

ITU SECRETARY-GENERAL MILI was the top brass at the dedication of United Nations' club station K2UN on October 21st. W1AW and 4U1ITU in Geneva joined in the inaugural ceremony, though K2UN's current antennas toward Europe leave much to be desired.

ARRL'S BICENTENNIAL CONTEST could become an annual affair, though under a different name. So many participants said they liked its format that League Communications Manager W1NJM has asked for comments.

AMATEUR NOVICE TRAINING PROGRAM currently has more than five times the enrollment it had last year — nearly 35,000 vs 6,000 in 1975. Unlike previous years, established classes continue growing with nearly zero dropout while newcomers appear after hearing about them from friends or on 27 MHz.

CARF WILL HAVE a headquarters office complete with Amateur station thanks to a grant of $10,000 to the Kingston (Ontario) Old Timers' Amateur Radio Association. The office and Amateur station (VE3VCA) will be located at 370 King Street, West, in Kingston, but the mailing address will remain Box 356.

CANADIAN CB WILL EXPAND to 40 channels following the U.S.'s lead, but the General Radio Service operators won't be able to use their new channels until April 1. 40 channel radios for the Canadian market must be tested to tighter specs, but they can't be submitted for testing until after January 1. Commercial users presently operating on the new frequencies have the option of staying put or applying for a new assignment.

Whether A U.S. CBer entering Canada with a shiny new 40 channel radio before April 1 would have problems remains to be seen.

SEVERAL THOUSAND CB LICENSES issued around the beginning of 1976 apparently never reached the applicants. Areas affected include Zip Codes beginning with 0 through 8, and the problems ran from December through April. New licenses for the affected blocks are being printed and mailed; however, any licensee receiving one as a duplicate should discard it.

AMATEUR RADIO RELATED articles have been showing up in non-Amateur Radio publications with increasing frequency, but those not written by Amateurs often suffer from distracting inaccuracies. Ham Radio's editorial staff volunteers its services in reviewing future Amateur Radio articles for any non-Amateur publication that wishes to take advantage of the offer.

DX receiver

for the hf bands

This design features four selectable front ends and excellent dynamic range to help you dig out the weak ones

Today's erratic propagation conditions together with the tremendous increase in the number of amateur stations make DX operating much more difficult than, say, ten or fifteen years ago. In my case, using an SB220 linear amplifier and a four-element quad antenna, I had fair response to directional CQ DX calls only to become frustrated by adjacent-channel overload or interference in my receiver by strong local or short-skip stations.

The answer to this problem is a receiver with characteristics such as those in the design described in this article. Tests made on the 20- and 40-meter front ends of this design produced the data shown in table 1. Note that the dynamic range is 100 dB for both test conditions on 40 and 20 meters.

receiver front-end design

Conflicting requirements and tradeoffs were necessary to obtain a receiver with a wide dynamic range. Recent articles, studies, and experiments indicated that balanced mixers, together with selective rf tunable filters ahead of the first active device, represented effective means for diminishing third- through fifth-order intermodulation (IMD) and cross-modulation products.

I tried hot-carrier diode mixers using discrete diodes (monolithic packages weren't available) in both ring and cross configurations. It was an unsuccessful attempt, since low input-output impedances created matching problems and these mixers seemed to have an insatiable hunger for oscillator power.

The overall receiver noise factor is interrelated with the receiver passband; and in the mathematical formula, one term shows that it is also inversely proportional to the mixer gain. With the diode mixer, the receiver lacked sensitivity because of a conversion loss of almost 9 dB.

Without discrediting the hot-carrier diode mixer, better results were obtained by using the Motorola C6050G double-balanced integrated circuit and also selected pairs of the RCA 40673 dual-gate mosfet.

Occasionally, at certain antenna bearings, the combination of galactic, ionospheric, atmospheric, and man-made noise is below predicted levels; therefore an "in-out" rf amplifier would be justified. For higher-frequency bands, extremely good sensitivity is possible by using a balanced amplifier as in some vhf receivers.

Theoretically, this kind of arrangement should have a better noise factor, as somewhat more than 50% of internally generated random noise would be rejected in the common-mode operation, being cancelled in the balanced output circuit. The rf amplifier becomes a nuisance when the bands become crowded with strong signals exceeding S9+40 dB levels. Receiver gain compression occurs, along with numerous intermodulation products, indicating that it's time to turn off the rf amplifier! This is accomplished in this design by means of miniature reed relays.

Another unwanted phenomenon in mixers is so-called "reciprocal mixing." This mixing is a direct consequence of oscillator noise modulation. When a large interfering signal appears at the mixer input, the signal will mix with oscillator noise and, although the interfering signal may be out of the i-f passband, the noise so produced will be within the i-f passband.

Reciprocal mixing is measured by the amount of noise introduced by a closely spaced interfering signal;

By Ovi Florea, WB2ZVU, 76 Whitson Road, Huntington Station, New York 11746

i.e., when the level of the interfering signal, expressed in dB above 1 μV and spaced 20 kHz away from the desired signal produces a reduction of signal-to-noise ratio by 10 dB.[1] Oscillator noise reduction is possible by a) careful elimination of unwanted spurious generation on other frequencies, b) good dc filtering, c) good stability, and d) reducing oscillator power and narrowing oscillator bandwidth. Recommended low-noise oscillators are those consisting of differential amplifier integrated circuits with balanced output circuitry.

Ultimately, reciprocal mixing reduces mixer dynamic range by raising its equivalent noise floor. As stated above, inherent oscillator noise modulation is also dependent on oscillator power; therefore, mixers working at high injection levels will be more affected.

table 1. Measured characteristics of the DX receiver 20- and 40-meter front ends.

receiver characteristic	40-meter band rf ampl off	40-meter band rf ampl on	20-meter band rf ampl off	20-meter band rf ampl on
Sensitivity for 10-dB s/n	0.5 μV (−119 dBm)	0.1 μV (−133 dBm)	0.2 μV (−127 dBm)	0.07 μV (−136 dBm)
third-order intercept point (dBm)	+23	+6	+15	+3
gain compression (dBm)	+10	−10	0	−15
dynamic range (dB)	100	100	100	100

Front end. Four separate front ends are selected by rotary switch S1 (**fig. 1**). In this design resistors in the signal path were kept to a minimum as they are noise devices. Despite the cumbersome appearance, this setup allows individual band optimization and eliminates

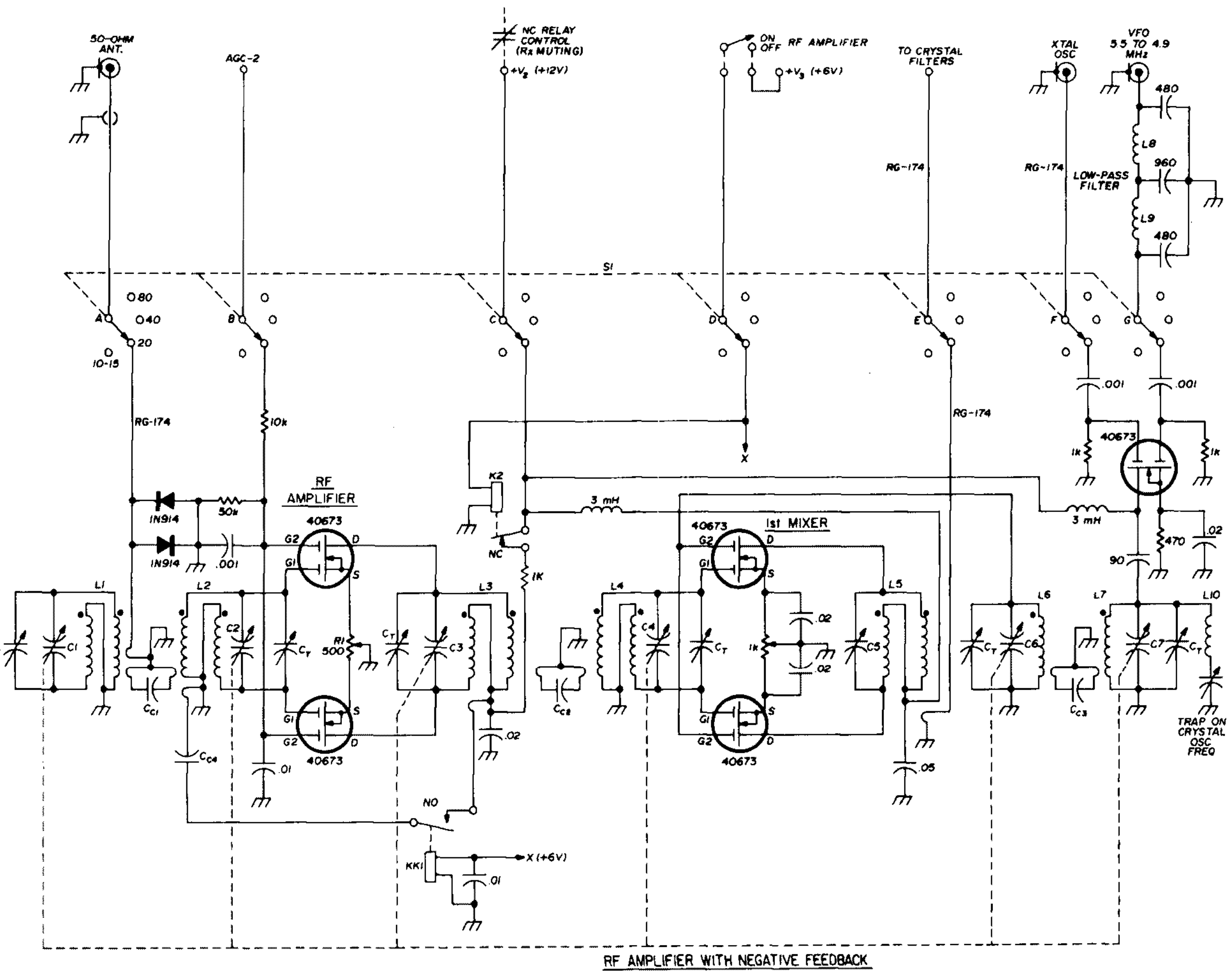
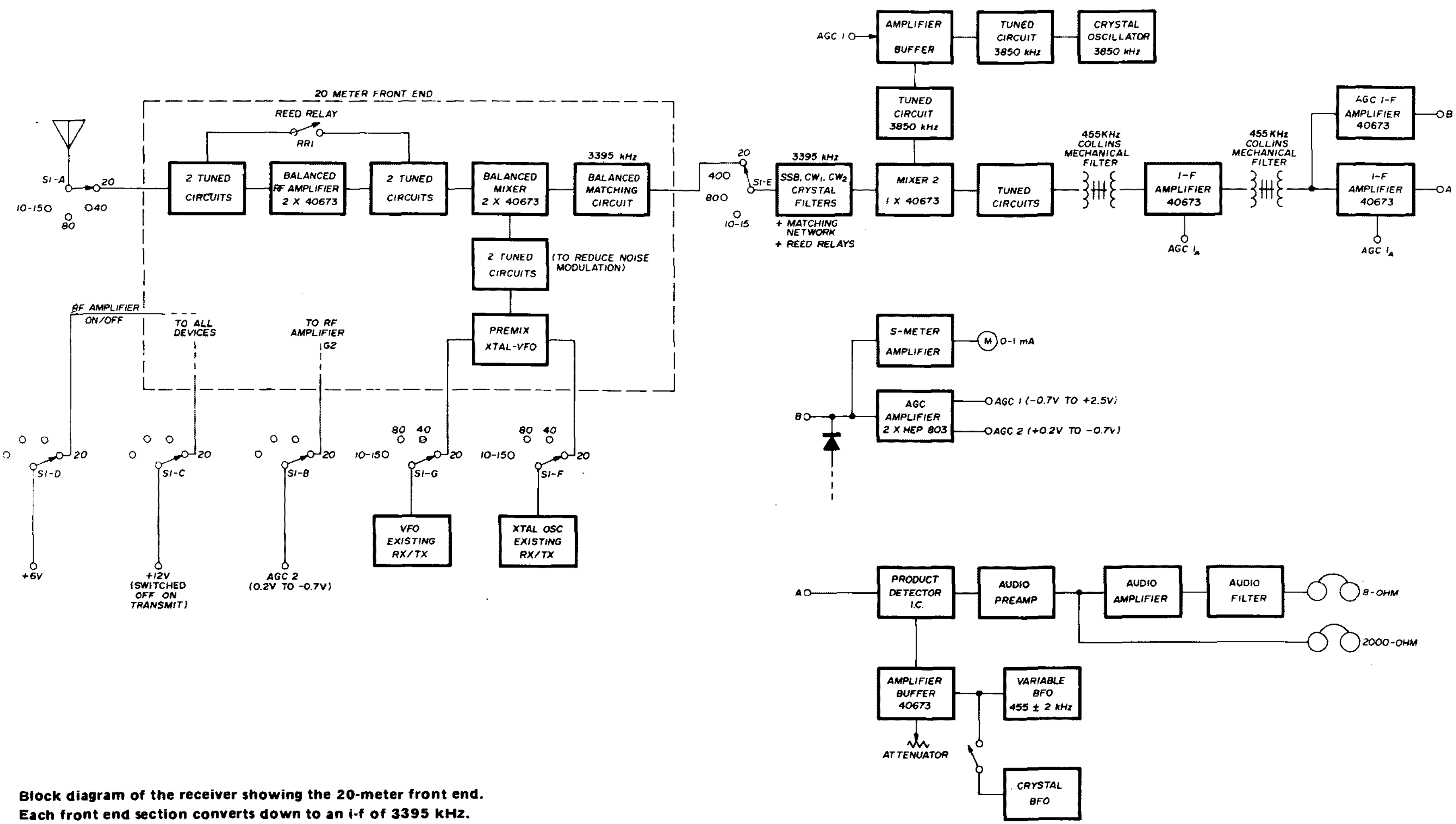


fig. 1 Schematic of the DX receiver front end, first mixer, and vfo. All links are one turn for an impedance of approximately 50 ohms. All trimmer capacitors are mica, 20 to 130 pF. L8, L9, 17 turns no. 24 AWG (0.5mm) on Amidon T-50-6 core. L5, 44 turns no. 28 AWG (0.3mm) wound 2 x 22 on Amidon T-50-6 core. The i-f is 3395 kHz.



Block diagram of the receiver showing the 20-meter front end. Each front end section converts down to an i-f of 3395 kHz.

problems related to balanced circuitry switching. I was fortunate enough to obtain some surplus vhf six-gang split-stator variable capacitors. To prospective builders of this front end, I suggest the use of smaller units of the kind having a shaft extension on both sides, which could be mechanically ganged. A good substitution would be a matched set of varactors for an unbalanced arrangement.

Good tracking between rf input and pre-mixer output circuits is imperative since the 4-pole rf filter is very selective, particularly on the lower bands. On the 20-meter band with the rf amplifier off, measured filter response was 50 kHz wide at 20-dB down, which means that an unwanted signal 25 kHz away from the receive frequency is attenuated 20 dB. The frequencies shown in table 2 would be applicable to Kenwood or Heathkit lines, but the principle could be extended to other makes.

Mixer. The dual-gate 40673 mosfet mixer has lower third- and fifth-order products in its output when gate G2 remains nonbiased; however, gain is definitely lower. For best performance oscillator injection level was set at 300 mV.

Fig. 2 shows the 40- and 80-meter alternative mixer configurations. The Motorola C6050G double-balanced differential amplifier with emitter degeneration has good linearity when the lower input (i.e., our oscillator port) is nonsaturated (maximum 27-mV oscillator level — see low values of CC3), while the upper quad could be saturated or unsaturated (our signal port). Recommended external biasing is:

$$(V6, V9) - (V7, V8) \geqslant 2\ V$$
$$(V7, V8) - (V1, V4) \geqslant 2.7\ V$$
$$(V1, V4) - (V5) \geqslant 2.7\ V$$

Filters. In accordance with modern receiver design practice, an ssb filter and two CW crystal filters are placed at the first mixer output through a matching network. Filter selection is accomplished by using remotely controlled miniature reed relays. Interelectrode capacitance of these reed relays is so minute that, when dc switching voltage is removed, signal feedthrough was measured at 120 dB down.

Fig. 3 shows the crystal filter arrangement. To obtain the above attenuation figure, filters were placed against a ground plane (double-clad printed circuit board, the face next to the filter grounded on one spot only). The second mixer is of classic design, and the local oscillator operates in a very stable mode. The second i-f strip is on 455 kHz and uses two Collins ssb mechanical filters.

Agc. An elaborate amplified agc system is derived from a separate uncontrolled i-f amplifier. One branch of this fast-rise, slow-decay agc system controls the i-f stages along with the local oscillator amplifier-buffer by

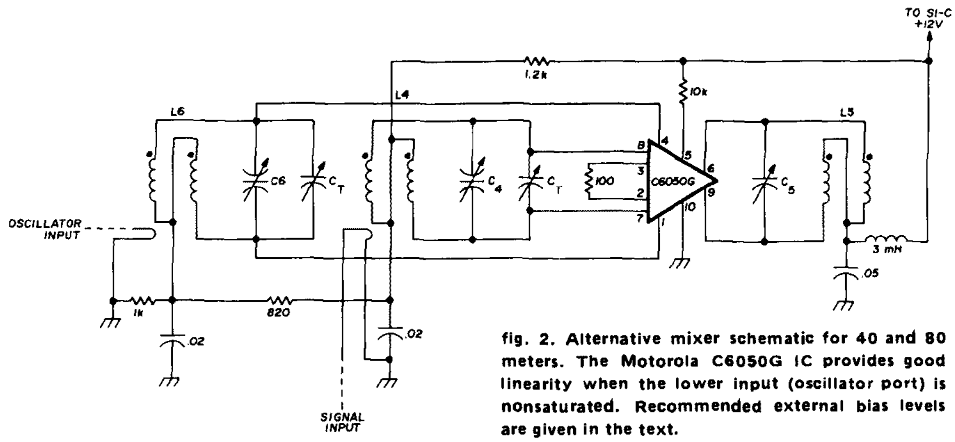


fig. 2. Alternative mixer schematic for 40 and 80 meters. The Motorola C6050G IC provides good linearity when the lower input (oscillator port) is nonsaturated. Recommended external bias levels are given in the text.

table 2. Coil and capacitor values for coverage between 3.5-28 MHz.

band (MHz)	frequency coverage (MHz) circuit 1 to circuit 4	circuits 6 and 7	vfo	xtal osc	coils and capacitors* L1 = L2 = L3 = L4 (turns)	L6 = L7 (turns)	C1 = C2 = C3 = C4 (pF)	C6 = C7 (pF)	$C_C1 = C_C2$ (pF)	C_C3 (pF)
3.5	3.5-4.0	6.895-7.395	5.5-4.9	12.395	2 x 20 T-50-6	2 x 16 T-50-6	10-30	8-20	800	90
7	6.9-7.9	10.295-11.295	5.5-4.9	15.895	2 x 14 T-50-6	2 x 12 L7 = 24T	10-30	8-20	470	51
14	13.9-14.9	17.295-18.295	5.5-4.9	22.895	2 x 7 T-50-6	12T T-50-10	10-30	8-20	350	330
21-28	20.5-30.0	23.895-33.395	5.5-4.9	29.895 36.895 37.395	2 x 5 T-50-10	8T T-50-10	10-100	8-90	250	220

*No rf amplifier on 3.5 MHz. Rf amplifier on 7 MHz is unbalanced

Notes:

455 kHz i-f transformer:

A&B —	primary impedance	Z_p = 35k	Radio Shack
	secondary impedance	Z_s = 6k	"black core"
C —	primary impedance	Z_p = 60k	"yellow core"
	secondary impedance	Z_s = 500k	
D —	primary impedance	Z_p = 40k	"white core"
	secondary impedance	Z_s = 1k	

SSB crystal filter Trio 4 pole 2.2 kHz at −6 dB
CW-1 crystal filter Trio 4 pole .5 kHz at −6 dB
CW-2 crystal filter Heathkit .4 kHz at −6 dB
(CF = 3395.4 kHz)

Reed relays 6V miniature for PC board mounting
(Electronic Applications Co. 1A3A-H)

All resistors 1/4 watt

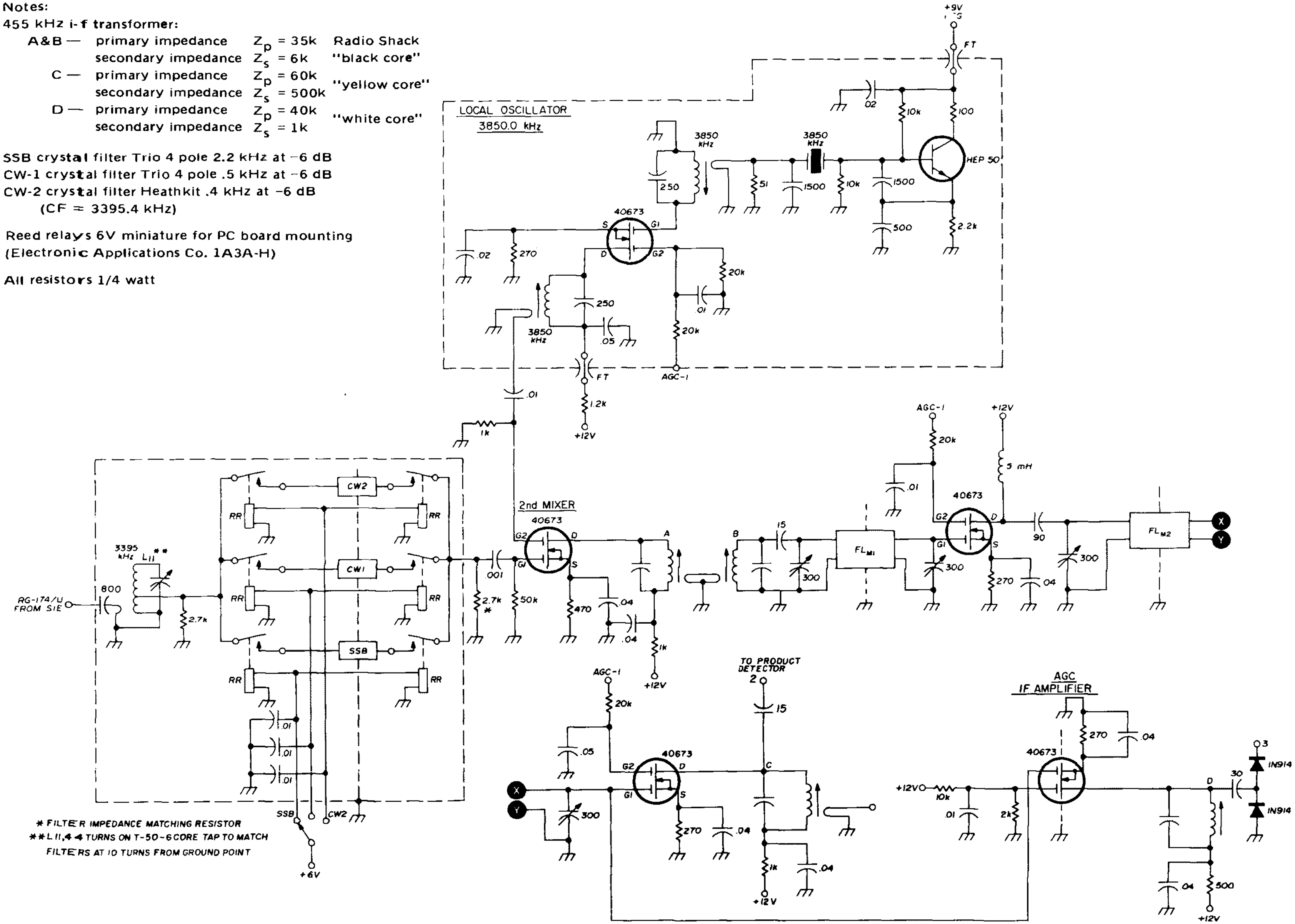


fig. 3. Crystal filter, local oscillator, second mixer, and agc amplifier schematic. Filter sections are selected by remotely controlled reed relays.

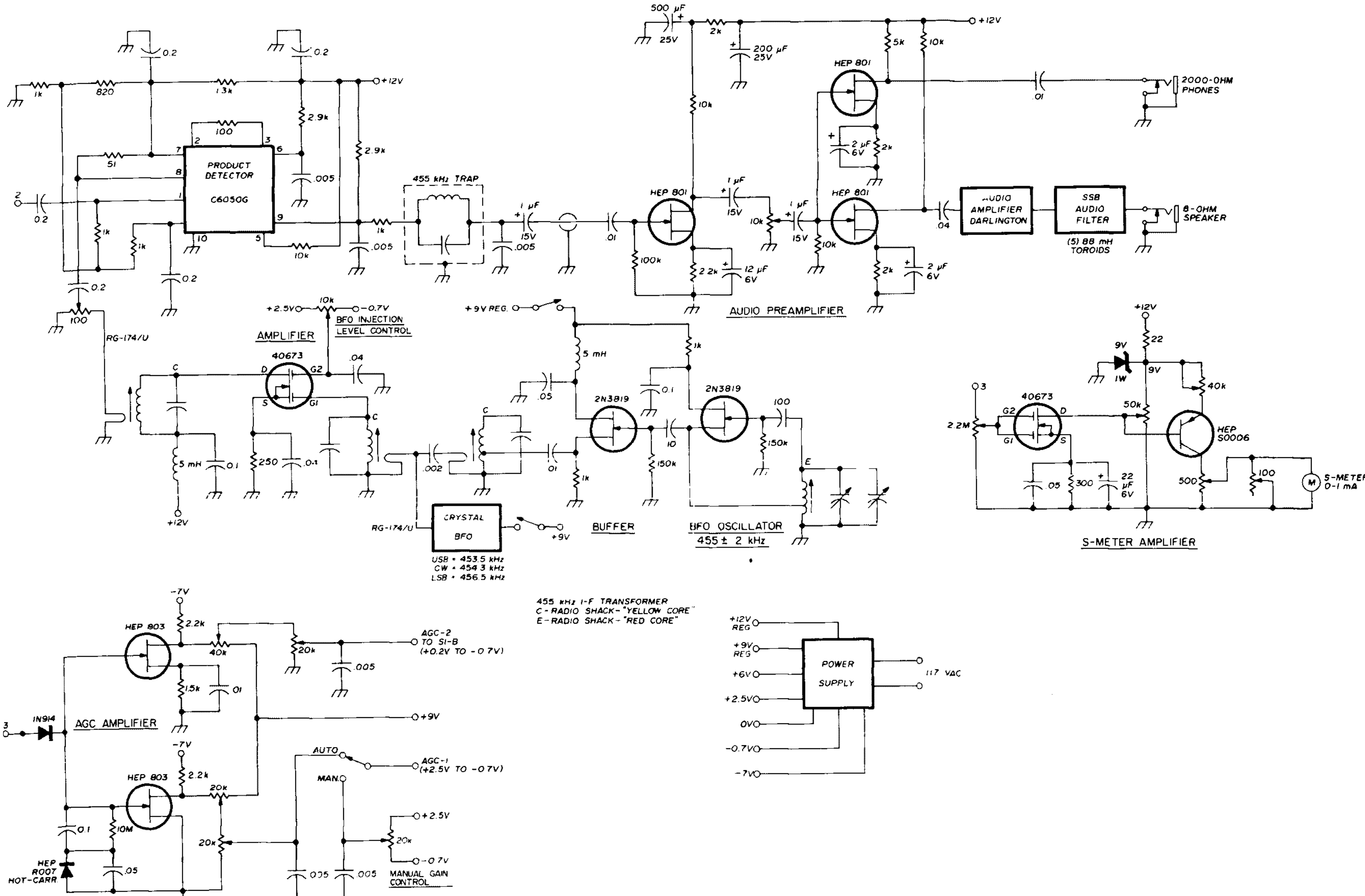


fig. 4. Product detector, bfo, agc, and audio circuits. The Motorola C6050G is used for the product detector, which produced excellent strong-signal rejection with decreased bfo injection voltage.

swinging voltages from +2.5 V (weak signals) to −0.5 V (strong signals). The second agc branch is threshhold-biased and applied only to the rf amplifier over a restricted range, so that front-end linearity is not impaired. The receiver has a separate S-meter amplifier as shown in fig. 4. Minipotentiometers are used to calibrate the meter. Accuracy is within 3 dB, or one-half S-point.

Detector and bfo. The product detector and bfo are of particular interest. A very sensitive Motorola C6050G double-balanced IC makes a good product detector and requires a maximum of 300 mV bfo injection for the strongest signals.

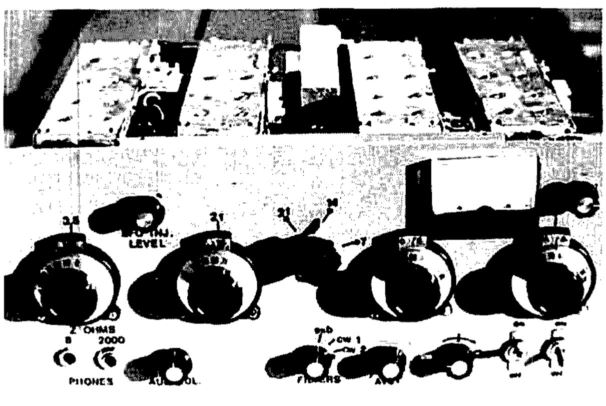
The beat-frequency oscillator is variable within 455 ±2 kHz. Its output is amplified and filtered to supply the required maximum of 300 mV into a 50-ohm load at the product detector.

Theory says that, for minimum distortion a product detector requires carrier injection at least 10 times the level of the incoming signal at the input port. What would happen upon lowering that injection level? Decreasing the bfo level caused strong perturbing signals to become unreadable, while weak DX signals were still crystal clear. One helpful finding was that the audio output versus bfo level dropped about 10 dB faster for the stronger signal compared to the weak one.

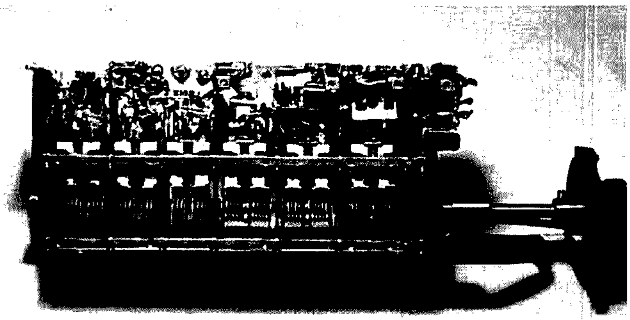
A crystal bfo was incorporated for good transceive operation. This crystal bfo is used to calibrate the variable bfo. At times, the output from the product detector is extremely low; henceforth, a noiseless audio preamplifier is required. The receiver also employs a 5-toroid audio filter to alleviate inherent noise pickup when using an audio amplifier.

construction hints

This is an experimental receiver; appearance was secondary to conveniences such as avoiding stray noise pickup, rf feedback, and audio hum. It was constructed by degrees starting with the front end for the 20-meter band. Mosfet devices were first tested with respect to dynamic transconductance and dc for matching purposes. Each device was set up in an amplifier with a plug-in socket and identical rf signal levels were applied



Front view of the DX receiver with cabinet removed.



Typical receiver front-end arrangement. Capacitor stator plates are easily removable so that proper maximum-minimum capacitance is obtained for band coverage and precise tracking.

to both gates from a signal generator, while a vtvm was used to measure output level on a tuned circuit. A table was compiled, and devices showing close characteristics were paired. From 40 available devices I could select only three closely matched pairs, which leads to the conclusion that integrated mosfet circuits would be a far better choice.

tests and results

The 20- and 40-meter front ends were tested for sensitivity, IMD, gain compression, and dynamic range (table 1). Two powerful signal generators (most signal generators are not capable of delivering more than 100 mV into a 50-ohm load) were used, tuned 10 kHz apart, each at a level of +12 dBm, which becomes +6 dBm after the hybrid combiner. The spectrum analyzer indicated at this level an IMD of 35 dB on the 40 meter band.

acknowledgement

I wish to express gratitude to Serge Costin, WB2ZWJ, for his continuous technical advice in the design and construction of this receiver.

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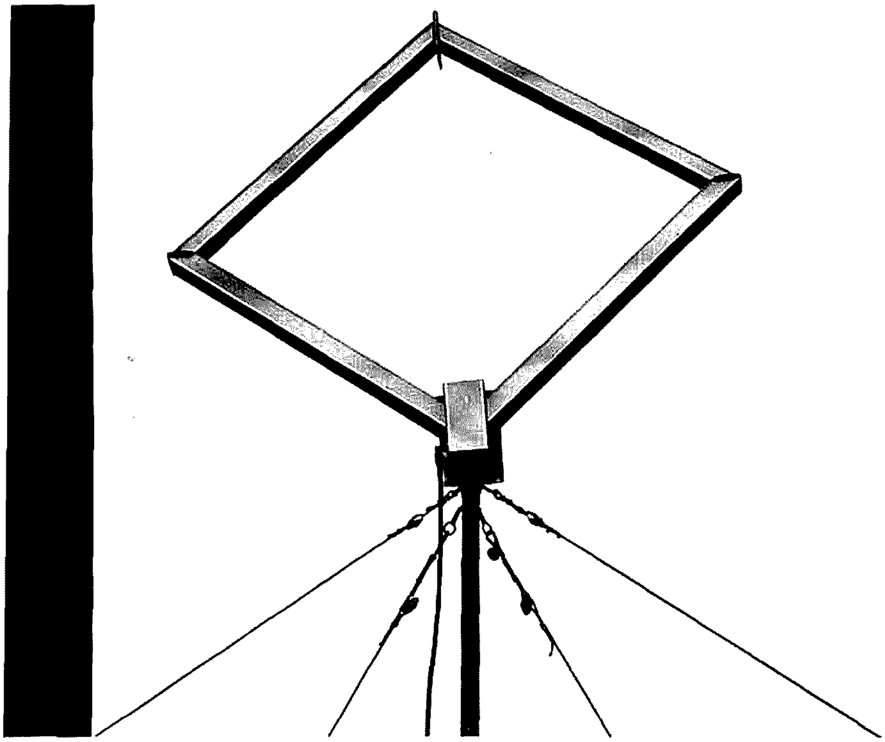


loop antennas

A discussion of small loop receiving antennas and details on their construction for the low-frequency bands

Most amateurs use the same antenna for both receiving and transmitting. This makes a lot of sense on vhf, and on 10, 15 and 20 meters, but at the lower frequencies which are more susceptible to noise interference entering via the antenna (160, 80 and 40 meters), the small loop receiving antenna has some advantages in reducing the susceptibility to certain types of noise. This article will attempt to expain the loop's operation in the simplest possible terms and will describe several practical loop antennas which are suitable for amateur use.

The electric- and magnetic-field components of an incoming electromagnetic wave are at right angles to each other. The plane formed by these components is at right angles to the direction of wave arrival. With the wave polarization and the direction of wave travel shown in **fig. 1**, both the electric and magnetic field components excite current flow in the vertical portions of the simple unshielded loop. The current induced by the electric field is due to the difference in charge impinging along the length of the vertical elements, while the current due to the magnetic field is because of the motor-generator action of the vertical conductors cutting the lines of force in the magnetic field as it moves past the conductors.

The currents due to both field components are mutually in phase, and although neither the electric nor the magnetic field components can exist without the other in the radiated electromagnetic field, the loop antenna behaves identically with excitation from either or both field components.[1]

While the voltage available at the terminals of a dipole is simply proportional to the current induced in the dipole, the voltage available at the terminals of a small loop is proportional to the *difference* between the

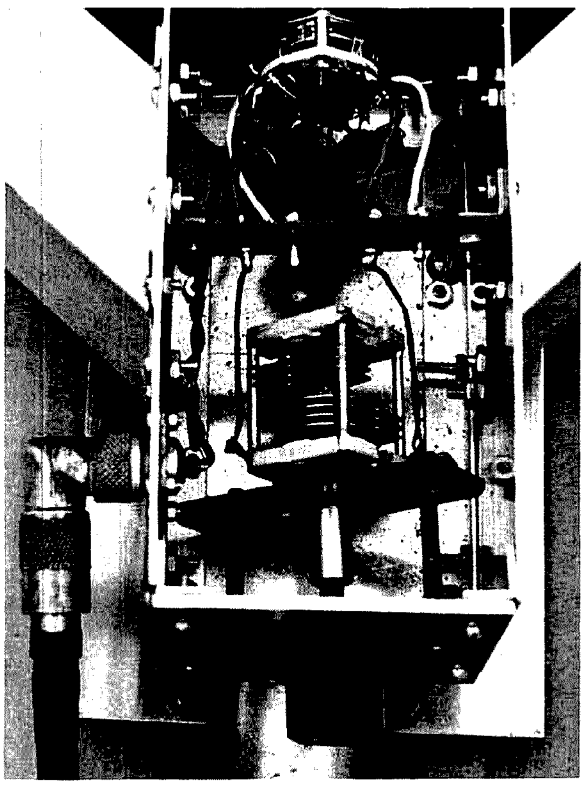
By John R. True, W4OQ, 10322 Georgetown Pike, Great Falls, Virginia 22066

currents induced in the two opposite vertical loop elements. No currents are *induced* in the top or bottom horizontal conductors connecting the vertical elements.

When the *axis* of the loop is pointing toward the signal source as in **fig. 1**A, the two vertical elements of the loop are excited at the same phase point of the wave front. Thus the current induced in both elements is of the same amplitude and phase, and flow in the same absolute direction (see **fig. 2**). However, the two currents are actually flowing in opposite directions with respect to a continuous, one-way travel around the loop, and therefore cancel each other, producing zero net voltage.

On the other hand, when the *plane* of the loop is pointing toward the signal source, as in **fig. 1B**, maximum voltage is produced because the two vertical elements are now in positions of maximum *difference* in phase relationship with the wave front, with the resulting difference between the currents induced in the vertical elements producing a maximum voltage. In fact **fig. 1B** shows that during the portion of the wave cycle when the field is changing most rapidly, the currents in the two elements are flowing in *opposite* absolute directions (one flowing upward and the other downward), with the result that both currents are actually flowing in the *same* direction around the loop, and are therefore mutually aiding instead of opposing as in **fig. 1**A. For orientations of the loop at angles in between the two just described, and in general, the voltage produced is proportional to the cosine of the angle formed between

Junction box for the square loop antenna containing 80/40 bandswitch, balun and tuning capacitor.



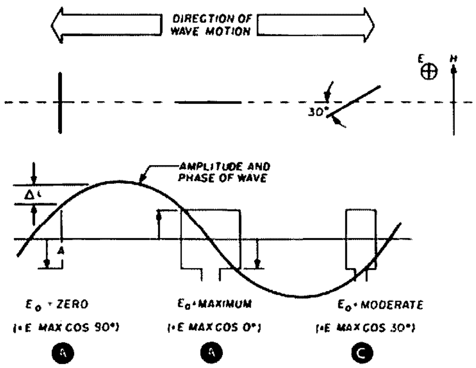


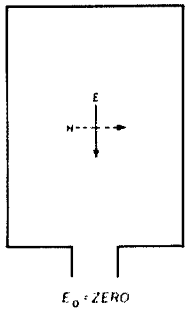
fig. 1. Directivity of the loop arises from the interception of the wavefront by each half of the antenna. In (A) the currents are of equal amplitude and phase, producing no net voltage difference at the output. In (B) the currents are at maximum amplitude and phase difference, producing maximum output voltage. In (C) the current difference between loop halves is moderate, producing a corresponding voltage difference at the output terminals. Voltage output is small for angles larger than 60 degrees.

the loop plane and direction of the wave propagation, as in **fig. 1C**. The resulting figure-eight radiation pattern is thus a perfect pair of circles tangent to one another as shown in **fig. 3**.

For several basic reasons, nearly all practical loop antennas are electrostatically shielded by means of an open-turn shield. One reason is that electrostatic shielding is a convenient way of achieving a capacitance balance between the two opposite halves of the loop and ground. Without this balance the figure-eight pattern would be distorted and the nulls misplaced and obscured. Second, the open-turn shield shown in **fig. 3** forms a balun, permitting the loop to feed an unbalanced load without upsetting the loop-to-ground balance. And third, electrostatic shielding renders the loop insensitive to the electric component of a passing wave. This has an insignificant effect on the reception of a wave propagated in the far field (radiation field). However, in the case of several types of man-made noise interference, the effect is to reduce the reception of the noise.

If the electrical disturbance producing the interfering noise is confined primarily to the *induction* field (as many such noise disturbances are), the electric compo-

fig. 2. Electric (E) and magnetic (H) components of a wavefront impinging upon a loop antenna, showing current flow in the loop. When the plane of the loop is parallel to the wavefront, as shown here, output voltage is minimum (see fig. 1A).

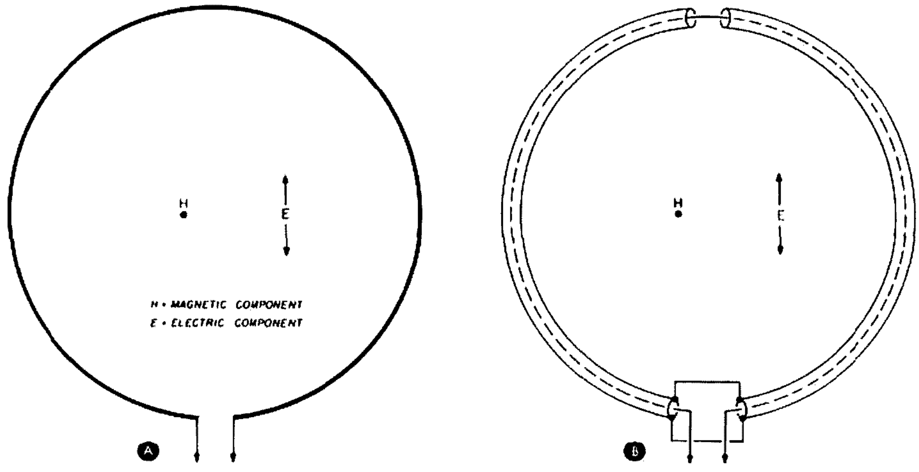


nent generally predominates over the magnetic field. Since the shielded loop is sensitive only to the magnetic field, there's a noticable reduction in noise pickup as compared to that of a vertical dipole. Providing the desired signal is not arriving from the same direction as the noise, some additional reduction in noise interference level is also available due to the directivity of the loop radiation pattern. Simply pointing the axis of the loop in the direction of the noise will minimize the noise pickup, while the desired signal still arrives from a favorable angle on the directivity pattern.

In general, atmospheric noise is propagated as a *radiation* field, generated by the electrical discharges that attend thunderstorms, both locally and throughout the world. Noise from an electrical storm concentrated in a single direction may be reduced by the directive properties of a shielded loop, but not by its insensitivity to the electric field. On the other hand, interference from precipitation static will be effectively reduced by the shielding properties of the loop because precipitation static is caused by an *induction* field localized directly around the receiving antenna.

The illustration of **fig. 1** is greatly exaggerated. When a loop 6½ feet (2m) between legs is used on 80 meters, the maximum wavefront intercepted represents only a small fraction of the energy intercepted by a half-wavelength dipole. However, such a small antenna is still adequate for good signal reception.

References 2 and 3 state that a maximum wire length of about 0.08 wavelength will produce currents throughout the loop of relatively equal amplitude and phase. This condition will produce the figure-eight pattern illustrated, but lengths in excess of this criteria will cause some pattern distortion. Reference 4 states that loops as large as 0.1 wavelength in diameter (0.314 wavelength circumference) can be used without serious pattern distortion. However, this reference is confined to aperiodic loops, while reference 2 deals with loops that are tuned, providing higher Q. The higher Q changes the current and phase difference in the loop wire, resulting



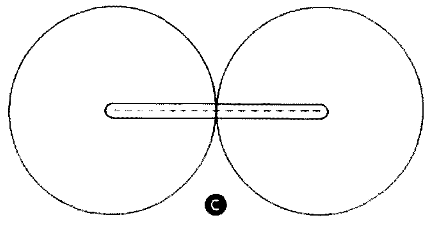
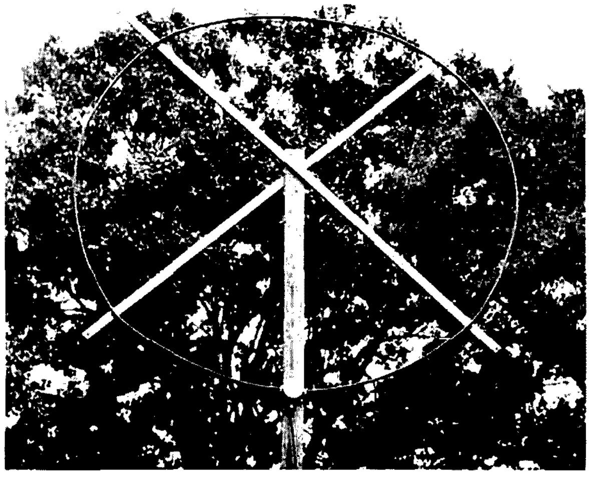


fig. 3. Unshielded loop antenna (A) and shielded loop antenna (B). There is no difference in the radiation patterns of the unshielded or shielded loop (C) if the output is balanced and if the capacitance to ground from both sides of the loop are equal. Loop pattern distortion results from these types of imbalances, and shielding the loop is one way of achieving the desired balance.

The coaxial loop of fig. 4 as built by the author.



in the shorter specified loop lengths (0.08 wavelength maximum).

The advantage of dual-band reception from a single loop further compromises the design criteria. For those who wish to retain the criteria of references 2 and 3, table 1 has been included. Corrections to table 2 must be made to compensate for this altered construction. Single-band designs would do well to follow the referenced design for maximum performance.

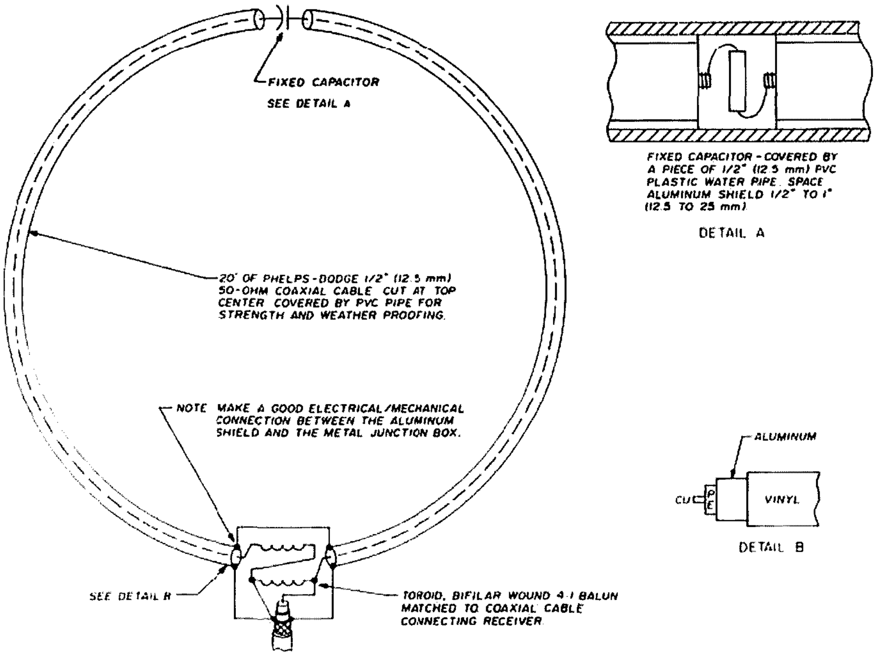


fig. 4. Single-turn coaxial loop antenna suitable for use on 80 and 160 meters. The toroid transformer balances the loop output to ground, whether shield is present or not. Typical dimensions are listed in table 2.

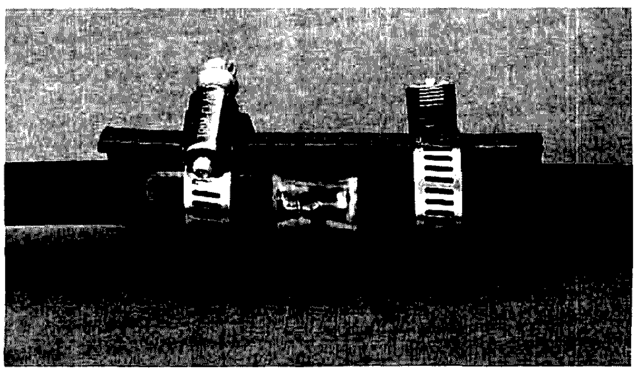
effects of nearby re-radiation

If there are metallic structures near the loop antenna of ample conductivity and size, such as power lines, homes with electrical wiring, water and furnace air-conditioning ducts and piping, as well as antennas, towers and guy wires, etc., signals will be injected into a loop antenna from both the direct signal and re-radiation from these nearby structures. If such structures have high Q and are at or near resonance at the frequency of the exciting signal, their energy may approach that of the direct signal and cause appreciable deviation to the true bearing of the signal source. The resultant voltage induced into the loop will be the vector sum of the amplitude and phase of the multiple sources. Since the amateur is not generally interested in obtaining accurate bearings of signal sources, such deviation is relatively unimportant. What is of prime importance is the amplitude of the desired signal and, secondarily, the depth of

table 1. Maximum wire length for direction-finding loops as specified in references 2 and 3 (0.08 wavelength).

frequency (MHz)	wavelength (meters)	maximum wire length
1.8	166.7	43' 7" (13.28m)
1.9	157.9	41' 4" (12.60m)
3.5	85.7	22' 6" (6.86m)
3.6	83.3	21' 10" (6.67m)
3.7	81.1	21' 4" (6.49m)
3.8	78.9	20' 9" (6.32m)
4.0	75.0	19' 8" (6.00m)
7.0	42.9	11' 3" (3.43m)
7.1	42.3	11' 1" (3.38m)
7.2	41.7	10' 11" (3.33m)
7.3	41.1	10' 9" (3.28m)

Detail of the break in the coax at top of loop. See detail B in fig. 4.

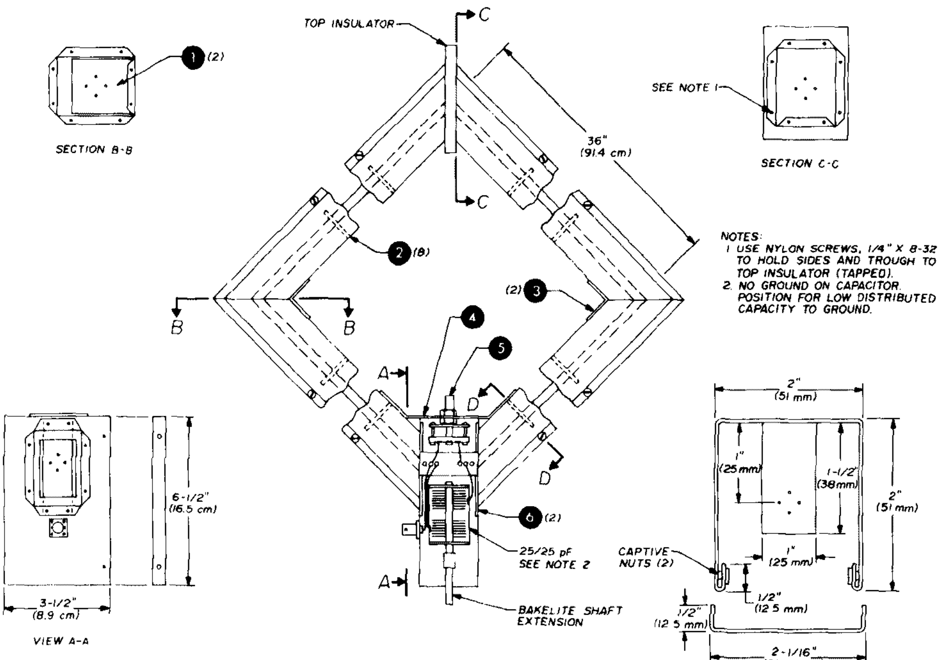


the null available to reduce the strength of an interfering signal.

If installed with the axis of the loop horizontal, only signals from the horizon i.e., low angle or ground wave, will produce the deep nulls shown. Signals from higher vertical angles will not have their wave front parallel to the plane of the loop and even though the azimuth is correctly set, the signal received will still be appreciable.* Therefore, do not expect many signals to show an extremely sharp null unless provision is made to tilt the loop axis in elevation as well as azimuth.

As shown in **fig. 4**, a practical loop antenna may be built from a single turn of coaxial cable. The shield must be broken as previously explained to remove the "shorted turn" effect. A loop so configured will almost completely shield the electric component of the wave. To insure retention of the figure-eight pattern the two halves of the loop must maintain symmetry as closely as possible.

Detail B of **fig. 4** shows the method used to insert a capacitor between the ends of the inner conductor as well as providing spacing of the outer shield. Any insulated covering that will provide strength to this point of the loop will be satisfactory, but it should also provide a weatherproof seal to keep moisture out of the break. Before placing the cover on the break, check for peak performance on your desired portion of the band. I used a grid dipper in the shack and tuned its signal in on the receiver. I noted each frequency for S-meter reading, then substituted various fixed capacitors (and combinations) to center the required bandpass.



item	description
1	corner insulator 2x2x1/16" (51x51x1.5mm)
2	insulator spacer, 1½x1x1/16" (38x25x1.5mm)
3	stiffeners, semi-hard aluminum, 2" (51mm) wide, 1/16" (1.5mm) thick
4	stiffener, semi-hard aluminum, 2" (51mm) wide, 1/16" (1.5mm) thick
5	five-pole, double-throw rotary switch
6	insulator, junction box, 1½x4x1/16" (38x102x1.5mm)

fig. 5. Double-turn, square-loop antenna for 40, 80 and 160. Typical loop dimensions and tuned-circuit components are listed in table 2.

An alternate construction method is shown in **fig. 5**. This illustration should provide most of the required construction details. The stiffeners at the junction box, J, and the two side corners were added to reduce the floppyness that existed without them. An even number of turns is required for symmetry since both the inductor and the capacitor are located in the junction box. Two turns (33 inches or 83.8cm on a side) are adequate for 80 and 160 meter operation. For 40 and 80 meters, two turns (16 inches or 40.6cm on a side) would comply with the design criteria of reference 1. For single-band operation, the lengths given in **table 2** provide an optimum signal-to-noise ratio and should result in maximum performance.

The tuneup procedure for the square loop is similar to that for the circular loop. Using a length of coaxial cable with a loop at the end attached to the output

*This feature of the loop enhances its ability to null out interfering signals, particularly local ground waves or electrical noise, while still maintaining reception of skywave signals. Editor

connector, the grid dipper is used to get the resonant frequency of the system near that required. (Caution: The grid dipper will also show a dip at the resonant length of the coaxial cable used.) Vary the number of turns on the primary of the toroid and the fixed paralleling capacitor until the loop shows peak pickup near your favorite spot in the band. Then position the loop where it is to be permanently mounted, with the required length of coaxial cable to connect it to the receiver, and inject a grid dipper signal into it. Plot the S-meter reading for each frequency. Adjust the inductor and capacitor until the bandpass is centered on the desired frequency.

I have substituted an alternate coupling network shown in **fig. 6** that provides slightly better pickup and higher loaded Q. It has been incorporated in all loops shown in **figs. 4, 5** and **6. Table 3** lists the required components.

The Hula-Hoop was separated, and the twinlead inserted. The loop is then spirally covered with folded (two layers, 3-inches or 76mm wide) heavy-duty aluminum foil with a number-15 (1.5mm) aluminum drain wire to provide connection for the foil shield. This was then continuously taped with vinyl electrical tape for weatherproofing. The aluminum wire is connected to the junction box to provide a path to ground for currents induced by the electric component of the wave.

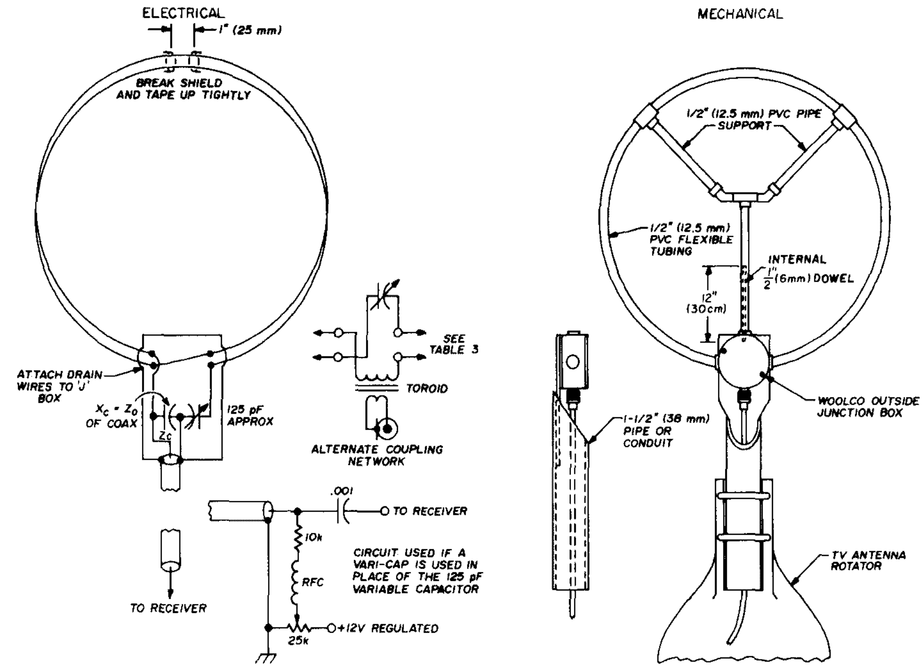


fig. 6. The Hula-Hoop designed by W5DS. The alternate coupling network is suggested. Component values for the coupling network are given in table 3.

table 2. Wire lengths and tuning capacitors used in the loops of figs. 3 and 4.

	frequency (MHz)	loop size	wire length	number of turns	toroid turns primary	toroid turns secondary	capacitor (pF) variable	capacitor (pF) fixed
coaxial cable loop	1.8 MHz	8'3" (2.5m)	26' (7.9m)	1	20/20 balun		0	500
	3.8 MHz	6'3" (1.9m)	20' (6.1m)	1	10/10 balun		0	125
square loop	1.8 MHz	36x36" (91x91cm)	48' (14.6m)	4	40	8	15-15	0
	3.5 MHz	36x36" (91x91cm)	24' (7.3m)	2	24	6	15-15	40
	7.0 MHz*	36x36" (91x91cm)	24' (7.3m)	2	12	7	15-15	75

*This loop built before reading references 2 and 3 so wire is considerably longer than 0.08 wavelength. Although it works fine, the design criteria of those references is recommended.

A third arrangement is shown in **fig. 6**. It was suggested by Bob Edlund, W5DS, who has named it the "Hula Hoop Loop" due to the basic material used to support the loop wires. A single-turn of TV twinlead is used to provide a two-turn loop when series connnected.

I used a 10-foot (3-meter) length of ½ inch (12.5mm) PVC tubing in place of the Hula-Hoop. This in turn is supported by ½ inch (12.5mm) PVC water pipe and fittings to form a Y-shaped support for the loop. A ½-inch (12.5mm) wooden dowel, boiled in beeswax, was

inserted into the vertical member of the Y support, terminating in the junction box to increase strength against wind torque at this point. The coupling networks shown will provide a good match between the loop and the coaxial line to the receiver.

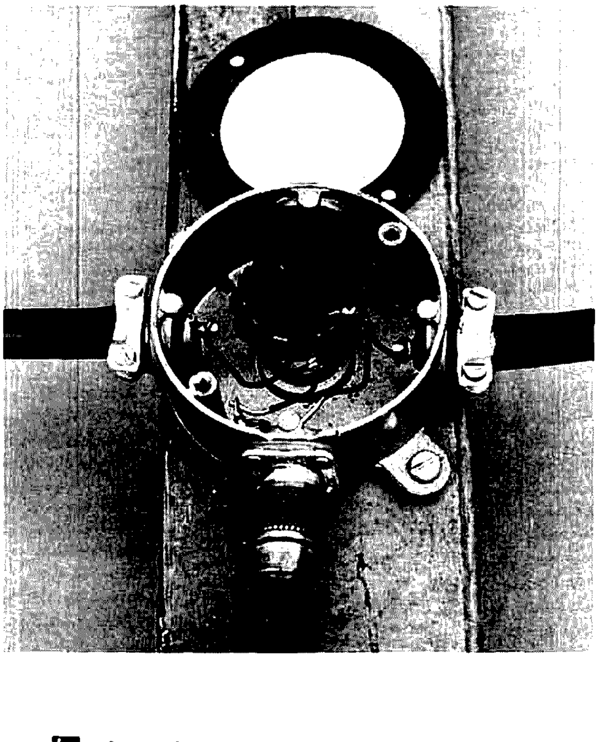
In an attempt to remotely tune the loop a voltage-variable capacitor (varicap) was substituted for the capacitor in the loop-coax coupling network. John Venters, K4UR, suggested, and provided, a silicon planar epitaxial diode (ITT type BA163) which, when reverse biased with 1 to 12 volts dc, provides a capacitance range of 10 to 260 pF. However, when feeding the varicap via the rf coaxial line with the required rf chokes to isolate the dc from the rf, the loaded Q dropped to about 10 (indicating the introduction of some form of undesired loss resistance). The convenience such a device would provide is worth further investigation; however, still to be found is a way to use the varicap and retain high Q. With the coupling networks shown, the loaded Q is in the vicinity of 75. This provides a 50-kHz bandwidth (at the 6 dB points) which is adequate on 80 meters if you operate near one spot most of the time.

table 3. Components for the alternate coupling network shown in fig. 5.

frequency	twinlead length	loop diameter	approximate capacitance (pF)	type	toroid primary turns	secondary turns
1.8 MHz	20' (6.1m)*	3'3" (99.1cm)	150	T106-2	14	8
3.5 MHz	10' (3.0m)	3'3" (99.1cm)	110	T68-2	12	7
7.0 MHz	5' (1.5m)	1'7" (48.2cm)	75	T56-2	10	6

*Two turns (be sure capacitor is centered in loop).

The 4:1 balun for the coaxial loop, mounted in the Woolco junction box.



In all loop configurations and couplers I tested there is a loss of about two S-units of signal pickup with respect to the vertical radiator I use for transmitting on 40 and 80 meters. Some of the weaker signals are then below the noise level of the receiver. A low-noise front-end preamplifier/preselector, similar to that described recently in *QST*[5] provides about 20 dB gain and puts the signal back up where the receiver can detect even the weakest signals.

conclusions

Comparing the loops against my "Five Band, Tower Antenna System"[6] for receiving, I get about five S-units reduction of man-made noise and precipitation static,* with only a loss of a couple of S-units of signal pickup.

Since the radiation resistance of such a small loop on the wavelengths involved is less than one ohm, it would make a very poor transmitting radiator.

*Atmospheric noise is propagated entirely by the radiation field so it *cannot* be reduced by using a shielded loop antenna. Precipitation static which is due to wind-blown rain, on the other hand, is an induction field and *can* be reduced by using a shielded loop. Editor

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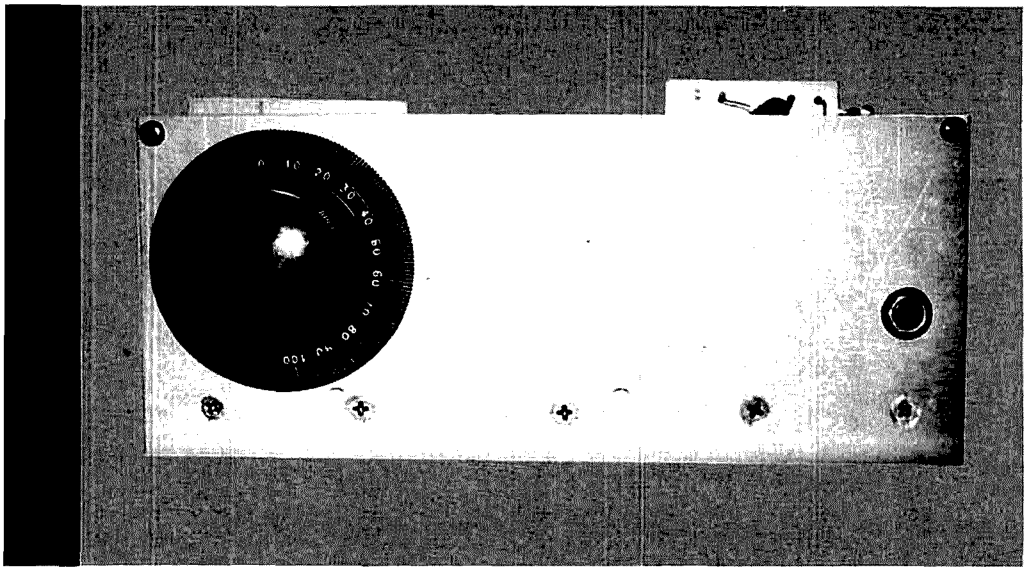


QRP transmitter

for 7-MHz CW

This little rig is still in the low-power class, but it will make a big difference on the crowded 40-meter band

The amateur radio literature over the past several years has described numerous solid-state QRP transmitters. Most of these transmitters have been in the power range of milliwatts, some a little higher and a few as high as 5 or 10 watts. These articles have made significant contributions to the development of better solid-state transmitters and have also kept interest high in building and testing such transmitters. The authors wish to thank all who have published their circuits and designs, thus enabling others to experiment and build designs of their own.

The CW transmitter described here is not just another low-power rig. It has some features that make it interesting, very inexpensive to build and, using a twenty-cent transistor in the final stage, it provides 20 to 30 watts input. The CW note is a pure delight to hear. All the transistors, which cost from twenty to fifty cents, are easily obtainable from Poly Paks, Radio Shack, and many surplus radio parts dealers.

The transmitter can be duplicated easily as it contains no critical circuits. It is built on five pieces of perf board, although etched PC boards would be the best construction approach. One such transmitter, built into a transceiver, has worked into most of the U.S., parts of Europe, and Australia.

power supplies

Before discussing the transmitter circuits, we'll talk about power requirements. A power supply must deliver

By Col. H. J. Stark, WA4MTH, 9231 Caribbean Boulevard, Miami, Florida 33189 and Dr. Si Marians, W4LPW, 6261 Collins Avenue, Miami Beach, Florida 33141

the quality and quantity of current and voltage required by the transmitter or problems will occur such as instability, poor performance, burned-out transistor, chirps, and key clicks.

This transmitter uses two separate power supplies. The first is a 12-volt, well-filtered and regulated supply capable of delivering up to 1 ampere. The transmitter will require about 250 mA from this supply. Reference 1 describes the power supply we used. The second supply powers the final amplifier transistor. It's a very useful power supply for transistor experimentation and test work, providing a well-regulated output to about 3 amperes at variable voltages between 5-30 volts. Its three transistors can be purchased for less than a dollar each. This power supply could be updated with an IC voltage regulator, but the circuit of reference 2 was used and is recommended.

circuit description

The transmitter has five stages, **fig. 1**. The vfo and doubler run continuously when the transmitter is keyed. The keyed stage, buffer amplifier Q5, Q6, operates in class C. Although collector voltage is present on this stage during operation, it has no rf output until the key is closed. Power to the vfo, doubler, buffer, amplifier, and driver is supplied by the 12-volt supply. The final

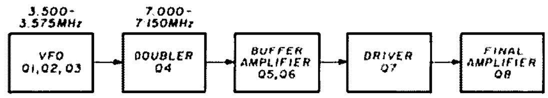
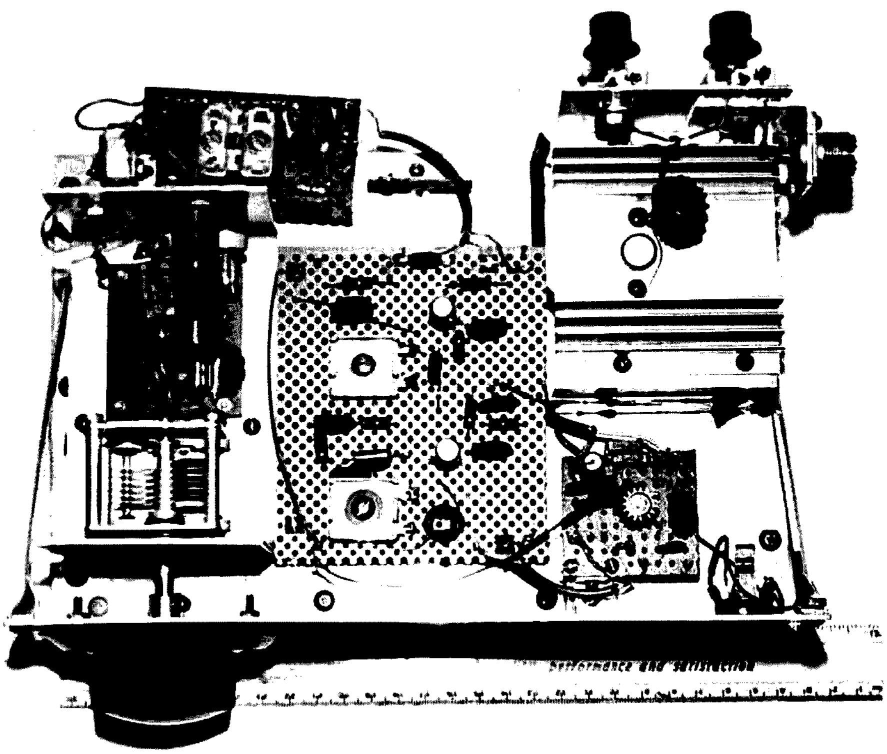


fig. 1. QRP transmitter block diagram. Eight low-cost transistors provide 2-30 watts input power for beefing up your CW signal on 40 meters.

amplifier, which also operates in class C, is powered by the variable 5 to 30 volt supply. Voltage to the other stages must be turned off during receive so you can hear the other fellow in your receiver.

Variable-frequency oscillator. The vfo (fig. 2) is straightforward, using an fet in a Colpitts configuration followed by an amplifier stage and an emitter follower for isolation. The vfo tunes 3.5-3.575 MHz. If you wish, you can make it tune the entire 75-meter band and have as much bandspread as desired. This circuit has little bandspread, however, as we're concerned with only 125 kHz of the 40-meter band, including the Novice portion.

The significant feature of this vfo is that frequency doubling from 3.5 to 7 MHz is used. This technique avoids pulling effects on the vfo frequency when keying the transmitter. The note you hear is the note that goes out over the air – crisp, clean, and chirpless. A 6.8-volt



The completed QRP 7 MHz CW transmitter. Vfo section is at left with the frequency doubler mounted at its rear. Buffer amplifier and doubler are shown at right, with the final amplifier removed from the chassis.

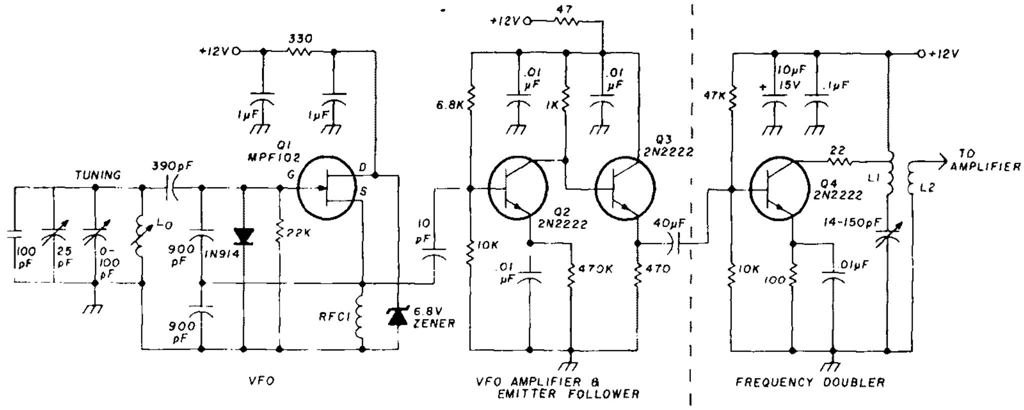


fig. 2. Vfo and frequency-doubler schematic. Circuit provides a stable, harmonic-free output to drive succeeding stages.

zener in the drain of Q1 and a high-speed switching diode in its gate ensure low harmonic content in the vfo output. The 1N914 acts as a clamp on the positive-going half cycle of the oscillator signal, thus preventing Q1 from reaching an operating point where high harmonic content may occur. Although many other transistors could have been used for the vfo amplifier, emitter-follower and doubler, the 2N2222 was chosen on the basis of performance and low cost.

Buffer amplifier. In this stage we begin to get a lift in rf output. **Fig. 3** shows the circuit, which uses two 2N3053s in class B. The 2N3053 will easily dissipate the power, provided it has a proper heatsink. The 12-volt supply to the collector must be well bypassed, and ferrite-bead chokes must be used on the collector leads. This is the keyed stage: the 2N3053 emitters aren't grounded until the key is closed. The component values shown in fig. 3 will permit this stage to function as a stable amplifier. Neutralization wasn't found to be necessary.

Driver. The driver is the simplest of all the stages. It operates as a class-C amplifier. The reliable 2N3053 is also used here, and it *must* be installed in a heatsink, otherwise it will be zapped. If this occurs, the 12-volt power supply will draw excessive current.

Final amplifier. So far we've been concerned with stages resonant at 3.5 or 7 MHz. In the final amplifier stage, however, we must think not only of having a 7-MHz resonant output circuit but must obtain an output impedance close to 50 ohms to match the antenna feedpoint impedance.

The final amplifier (**fig. 3**) also operates in class C. The input circuit to Q8's base uses a tuned-T network, and the collector output circuit uses a pi network so that the amplifier output impedance will be near 50 ohms. Rf

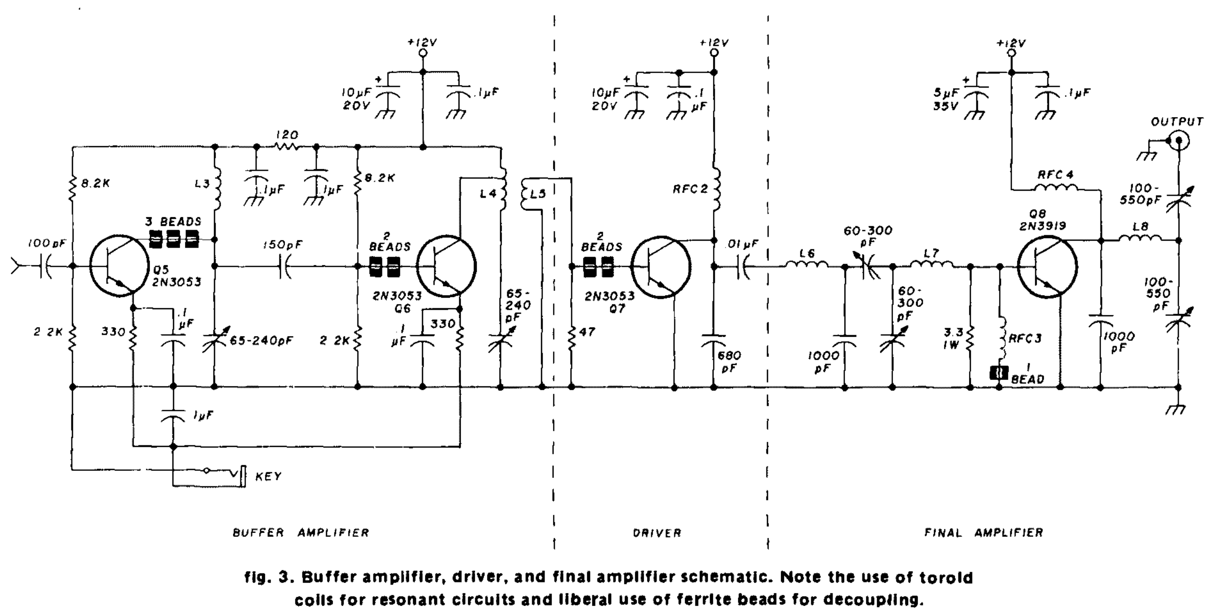


fig. 3. Buffer amplifier, driver, and final amplifier schematic. Note the use of toroid coils for resonant circuits and liberal use of ferrite beads for decoupling.

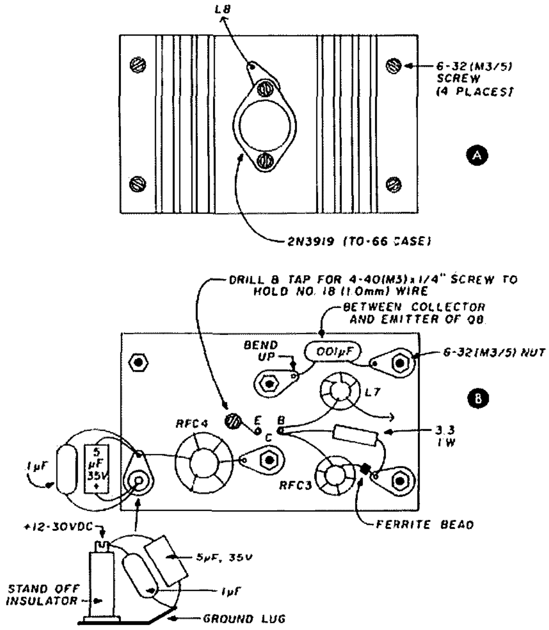


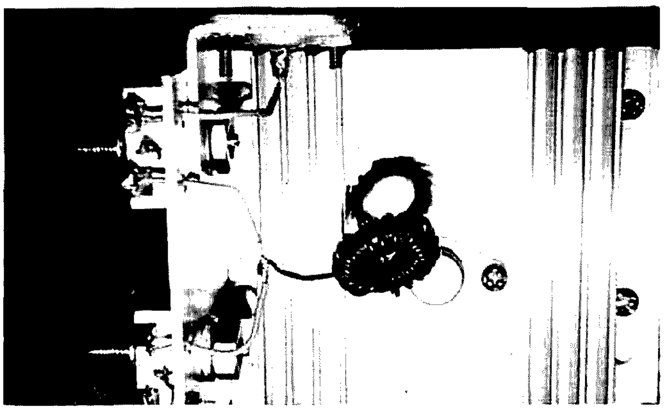
fig. 4. Construction details for the final amplifier. Sketch A shows mounting of the 2N3919 in its heatsink; B shows under-chassis wiring and parts location.

bypassing and rf chokes are important. This stage is built on a heavy aluminum heatsink, so the transistor barely gets warm while dissipating nearly 30 watts. It's hard to find a low-cost transistor on today's market to beat this one (Fairchild 2N3919 or Poly Paks 92CU1234). One word of caution: when testing you *must* use a 50-ohm dummy load. Adequate drive must be provided to this stage before it will function. References 3 and 4 are suggested for further reading.

construction

As mentioned earlier, the transmitter is built on five perf boards. Prepunched boards may be obtained from

Close-up view of the final amplifier. L8 is shown wound on the toroid.



such as Vector T-28 or flea clips (Lafayette Radio 198302) work well on this board. Arrange the parts on the board for a symmetrical layout, bearing in mind the necessity for short rf leads and adequate bypassing. Avoid crowding and try to keep dc voltage wiring isolated physically from circuits carrying rf voltages. Install a wire around the board periphery to serve as a ground plane for connecting all circuits to be grounded; this will avoid ground loops and feedback problems. The nice thing about perf-board construction is that you can rearrange parts until you're satisfied with the layout without touching a soldering iron.

table 1. Construction data for the inductors used in the QRP transmitter.

RFC 1	1 mH (not critical)
RFC 2	60 or 70 turns no. 28 (0.3mm) enamelled wire on T50-2 toroid core (Amidon). Fill core with wire.
RFC 3	25 turns no. 28 (0.3mm) enamelled wire on T50-2 toroid core (Amidon)
RFC 4	75 turns no. 28 (0.3mm) enamelled wire on T68-2 toroid core (Amidon)
L0	40 turns no. 26 (0.4mm) enamelled wire close wound on ½ in. (12.5mm) O.D. slug-tuned coil form
L1	27 turns no. 28 (0.3mm) enamelled wire on T50-2 toroid core (Amidon), tapped 7 turns from plus end of coil
L2	3 turns no. 22 (0.6mm) enamelled wire over L1
L3	21 turns no. 28 (0.3mm) enamelled wire on T50-2 toroid core (Amidon)
L4	27 turns no. 28 (0.3mm) enamelled wire on T50-2 toroid form (Amidon), tapped 7 turns from plus end of coil
L5	3 turns no. 22 (0.6mm) enamelled wire over L4
L6	14 turns no. 22 (0.6mm) enamelled wire on T50-2 toroid core (Amidon)
L7	14 turns no. 22 (0.6mm) enamelled wire on T50-2 toroid core (Amidon)
L8	17 turns no. 22 (0.6mm) or no. 18 (1.0mm) enamelled wire on T68-2 toroid core (Amidon)

Plastic transistor sockets are used for all devices except the final-amplifier transistor, which is packaged in a TO-66 case and mounted directly on a heavy heatsink. Coil data is provided in table 1. Fig. 4 shows parts layout for the final-amplifier stage. Note that this stage is built on a separate chassis, which includes the heatsink for the 2N3919 (fig. 5). The sketches in fig. 5 also provide overall dimensions for the frame that contains the five chassis.

test equipment

A few essential instruments will be needed for testing and making measurements. A vtvm with an rf probe capable of measuring up to 25-30 volts will be needed. The Heath Company stocks one that is a good buy, or if you want to build one a description appears in reference 5. Another useful item is the absorption frequency meter, also described in reference 5. A grid-dip meter is needed to determine resonant frequencies of toroid coils

and LC circuits. To save headaches later on, all toroid LC circuits should be tested before wiring them into the stages to ensure they are resonant and cover the proper frequencies.

A dummy load capable of handling about 8 watts is needed. It can be made by paralleling four 200-ohm, 2-watt resistors or some other combination of resistors

can in one or two more steps obtain an input of 25-30 volts at 700-850 or more mA. Thus at 30 volts at 850 mA you are putting 25.5 watts into the final transistor and will obtain 8 or more watts output. You can calculate your output power from E^2/R by measuring the rf voltage across the 50-ohm dummy load – pretty good performance for a few inexpensive transistors.

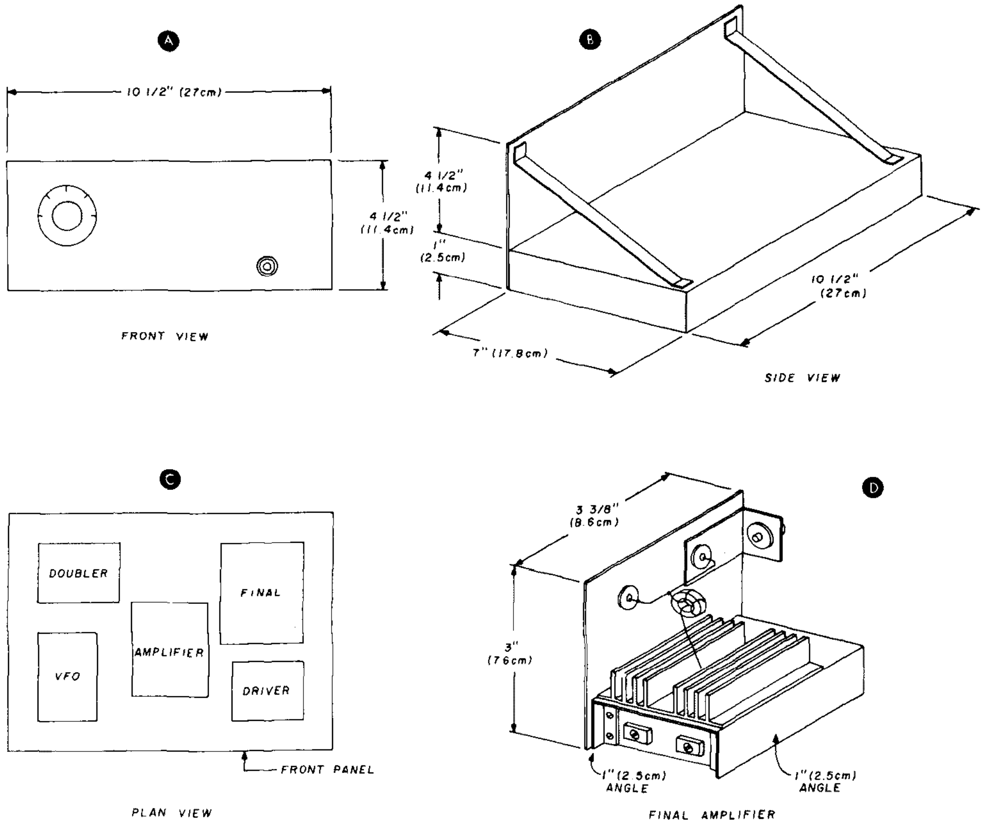


fig. 5. Dimensions and layout of the QRP transmitter. Sketches A through C show details of the main chassis; D shows the final amplifier arrangement.

to give 50 ohms and a power capacity of 8 watts or more.

When you complete construction of a stage, it should be tested with respect to the preceding stage. When you reach the last stage, be sure it has adequate rf drive. Connect the 8-watt dummy load to the SO239 output connector. You are now ready to test the completed transmitter.

Starting with 12 volts on the final-amplifier transistor, key the transmitter and tune all stages for maximum rf output using the vtvm probe. Two precautions: be sure your rf probe will handle 30 volts and *don't* test without dummy load. It's best to peak the final amplifier stage first, work back to the vfo, then repeak the final stage. Now you can release the key and advance the final-amplifier voltage to 20 volts. Key the transmitter and peak the four capacitors in the final. By working incrementally and peaking the final stage each time, you

When you connect the transmitter to the antenna, all you'll have to do is peak the capacitors in the final pi network for a maximum output. A tuning indicator is required here, such as an swr indicator. If there is a high swr, better antenna efficiency will be obtained with an antenna tuner.

references

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3. "Matching Network Design and Computer Solutions," *Motorola Application Note AN267.*
4. Wes Hayward, W7ZOI, "Increased Power for Solid-State Transmitters," *QST*, May, 1972, page 19.
5. "Rf Probe for Electronic Voltmeters," *The Radio Amateur's Handbook*, ARRL, 1975, pages 535-536.

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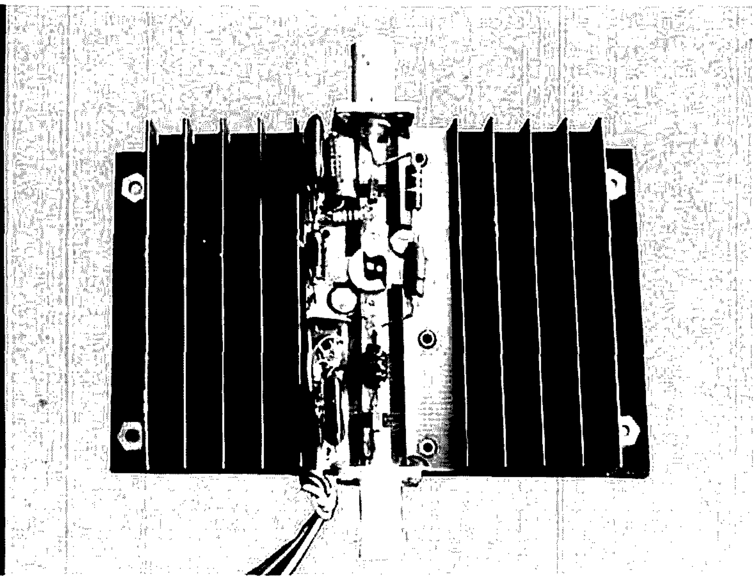


mospower fet as a broadband amplifier

A new type of power transistor provides linear performance over three vhf bands without switching

The *Mospower* fet introduced by Siliconix* in late spring, 1976, is undoubtedly the most revolutionary semiconductor in decades and one that will open up exciting new applications heretofore impossible with bipolar transistors. Switching of 1-ampere in less than 4 nanoseconds is commonly accomplished with this new power mosfet. Among the many new features is one in particular that will interest those who seek wide dynamic range: a linear transfer characteristic! Imagine too a transistor that can double for either a linear power amplifier or a wide dynamic range, low noise, small-signal, front-end transistor!

Other features of the *Mospower* device are typical for field-effect transistors and would be especially desirable for bipolar transistors. As with all fets, there is no thermal runaway nor secondary breakdown, and no minority-carrier storage time. The latter opens up interesting applications for class-D (switching) amplifiers. Additionally, the *Mospower* fet can accept any vswr – open or short – at any phase angle without debilitating effects. This enhancement mode, N-channel, mosfet can be operated in any class (A, AB, B, or C) without requiring a bias supply.

Unlike the usual mosfet, which is *planar* in construction, the *Mospower* fet is a four-layered vertical structure shown in fig. 1. This drawing compares, somewhat oversimplified, the fundamental differences between MOS, DMOS and VMOS – which is the generic name for the *Mospower* structure. Common to both MOS and DMOS (but not VMOS) is the singular disadvantage which affects their power handling capabilities: the geometry requires massive area to handle the necessary current. A further disadvantage of MOS and DMOS lies in their inability to accept high voltages. In the *Mospower* fet the current travels vertically, the source being on top while the drain is the bottom of the chip. In this vertical structure there are four layers whose dimensions are controlled by diffusion processes rather than by the less precise photolithographic methods common to planar (MOS) technology.

VMOS construction offers high current densities, high source-to-drain breakdown capability, and low gate-to-drain feedback capacitance, which makes the *Mospower* fet a great device for hf and vhf applications. Probably the most attractive aspect of this revolutionary semiconductor is its inherent linear transfer characteristic. In conventional mosfet (and jfet) devices this transfer characteristic is closely identifiable to a square-law

*Mospower is a trademark of Siliconix, Incorporated.

By Ed Oxner, Siliconix, Inc., 2201 Laurelwood Road, Santa Clara, California 95054

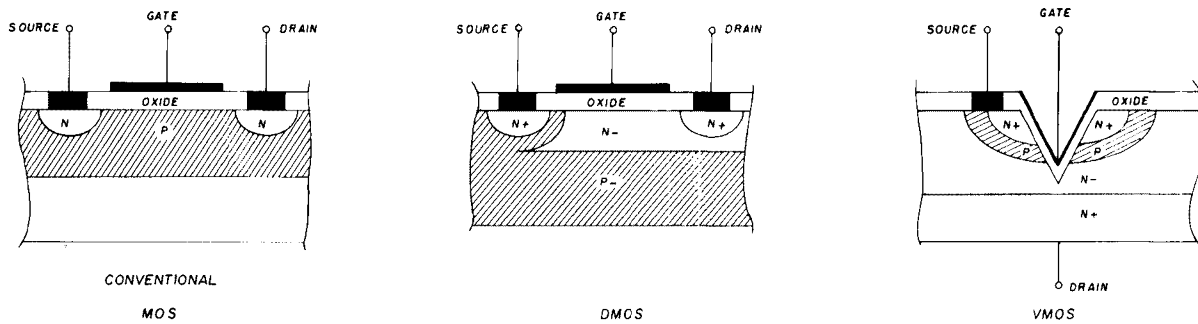


fig. 1. Fundamental differences in enhancement-mode MOS structures.

response; that is, the drain current is proportional to the square of the gate-to-source voltage. However, in the VMOS structure the short channel causes the drain current to be linearly proportional to the gate-to-source voltage. Fig. 2 is a transfer characteristic which shows this effect.

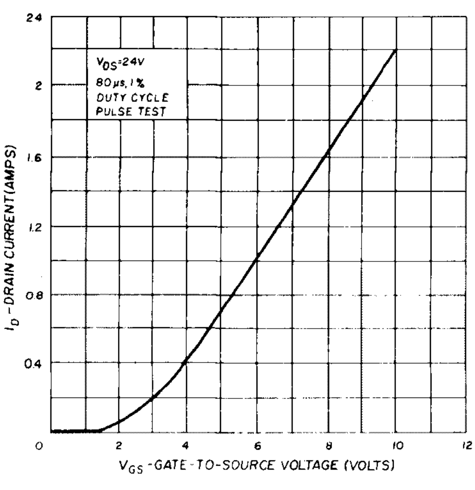
mospower vhf fet

The Siliconix VMP4 is packaged in the popular flange-mount, "opposing-emitter" (in this case, opposing-source) stripline configuration. This transistor is capable of saturated output power approaching 20 watts at 160 MHz. The performance shown in fig. 3 represents the available *saturated* output power versus frequency when both the input and output impedances of the VMP4 are conjugately matched (but *not* in the circuit described in this article). The input and output impedances (in the common-source configuration) are particularly well-suited for wideband amplifier service with *complete stability*. Unlike bipolar power semiconductors these impedances are affected very little by drive levels.

the circuit

Simplicity is an understatement for the wideband power amplifier shown schematically in fig. 4 and in the photo of the finished amplifier. Unlike many claims for broadband performance, this amplifier, by virtue of a negative feedback circuit, performs with flat gain response (±0.5 dB) over its entire operational bandwidth.

fig. 2. Transfer characteristic of the VMOS structure.



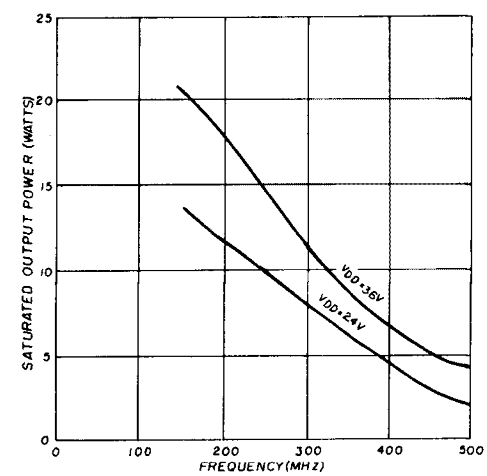


fig. 3. Showing the output power (saturated) versus frequency of the Siliconix VMP4. This assumes that the input and output impedances are matched. Drive power is 1 watt, drain current is 0.8 amp.

Two interesting features are immediately apparent in the circuit diagrams: first, the simple 4:1 transformer for broadband input matching and, second, effectively no matching circuit at the output. My philosophy is, "Why use parts if they're really not necessary?" The drain circuit needs no further complication. Some readers may question the wisdom of such an over-simplistic design, especially in light of the familiar equation

$$R_o = \frac{(Vcc - Vsat)^2}{2P} \qquad (1)$$

where R_o = output impedance, V_{cc} = drain supply voltage, V_{sat} = saturated drain-to-source voltage, and P = output power.

However, using this formula and making a few first-order assumptions, we can arrive at some near 50-ohm values

for the drain load impedance; for example

$$R_o = \frac{(25 - 3)^2}{2 \times 4} = 60.5 \text{ ohms}$$

To reach the lowest operational frequency requires a ferrite core with reasonably high permeability, but in this design 6 meters was my low-end goal and the 220-MHz band uppermost, so an operational bandwidth extending from 40 to 265 MHz was chosen. Only one circuit trick was required to reach the upper frequency objective and in reality it wasn't so much a circuit trick as a careful selection of component. I do not recommend the use of a commercially available molded 0.15 µH feedback inductor – manufactured inductors appear to exhibit too much distributed capacitance for this application. I used 6 to 8 turns of no. 30 AWG (0.25mm) enameled wire on a ½-watt, 1-megohm resistor. If you have an inductance bridge you can wind the choke to 0.15 µH, otherwise you may need to experiment. Using a commercially molded choke will severely reduce the upper-frequency limit.

About the only difficult aspect in building this circuit is preparing the double copper-clad board to accept the flange-mounted stripline transistor. Careful layout and cutting is required. As with any rf layout be sure to connect both copper foils (top and bottom) together, either with small eyelets or what-have-you. Additionally, remember that the *Mospower* fet is MOS and has an unprotected gate, so don't handle it without first being absolutely sure that you are not carrying a static charge. Stay off rugs and out of crepe-sole shoes until you've got the transistor soldered into your circuit. Once in the circuit you're free to do anything you want to with your amplifier.

Since this transistor operates with healthy currents it is absolutely necessary to mount the flange to a suitable heatsink. The large one shown in the photo is an over-design but it does emphasize that a heatsink is necessary. As is typical with any heatsink, you should use a suitable silicon grease or thermal compound between the flange and the sink.

A second precaution common to any high-current load is to avoid current-carrying molded chokes that may vaporize when the power is applied. I have found, quite by accident, that generally speaking, values of inductance *less than* 0.22 µH will hold up with currents of 1.5 amp or less. Further proof of reliability in regard to this construction is that I have built four identical amplifiers and all performed equally well.

An interesting aspect of this wideband amplifier is that performance does not seem dependent upon whether you use it for a small-signal amplifier, say in the microwatt area of a front-end receiver design, or for medium power (1 to 2 watts) amplification, possibly to excite a linear final amplifier.

Of special interest to those advocates of wideband amplifiers will be the observation that in a *Mospower* fet amplifier the wideband noise is literally unmeasurably low! For an example, this VMP4 (and any other VMP device) offers excellent small-signal noise figures. A typical value of 2.4 dB at 146 MHz is easy to achieve

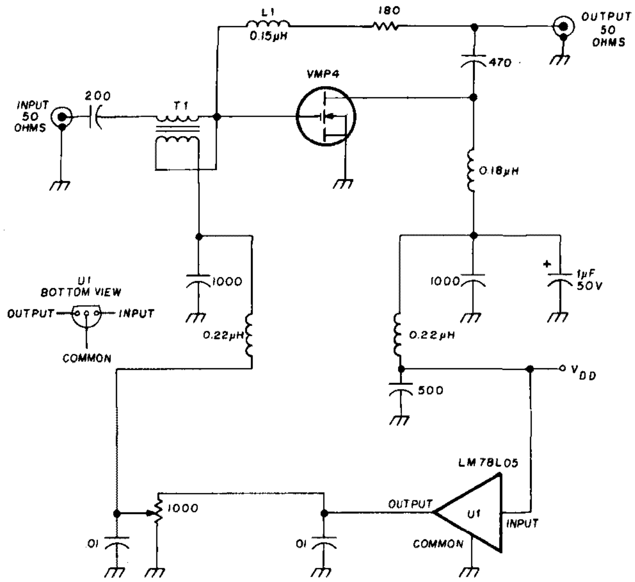


fig. 4. Simplicity is the keynote in this broadband amplifier. L1 should not be a molded choke; see text for winding information. T1 is 4 turns no. 22 AWG (0.6mm) twisted pair on an Indiana General F625-9Q2 toroid core.

with a properly matched input circuit. However, it should be pointed out that this particular circuit using the 4:1 matching transformer is *not* properly matched for optimum small-signal noise figure; that was not the objective in the first place. Bandwidth for two power levels is shown in fig. 5. With 1-dB compression occurring at an input level of +23 dBm, the +27 dBm input level is understandable under compression, hence the lower gain figure.

Two-tone, third-order intermodulation performance at both the 100-mW and the 1-watt output levels is displayed in fig. 6 as intercept point (IP). This point was calculated with reference to a single-tone output, using the formula

$$IP\,(dB) = P_{out}\,(dB) + \frac{P_{imd}\,(dB)}{2} \qquad \textbf{(2)}$$

When calculating the intercept point, or when comparing specifications between devices, care must be

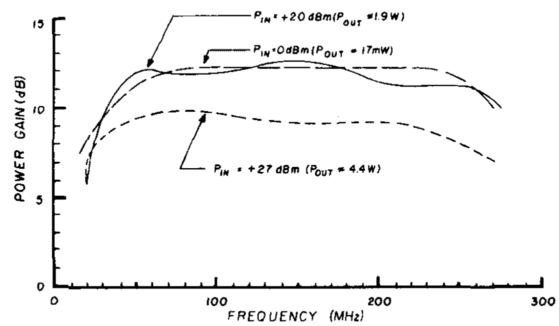


fig. 5. Power gain versus frequency curves at different levels of drive. The +27 dBm curve is above the compression point, therefore the gain is lower.

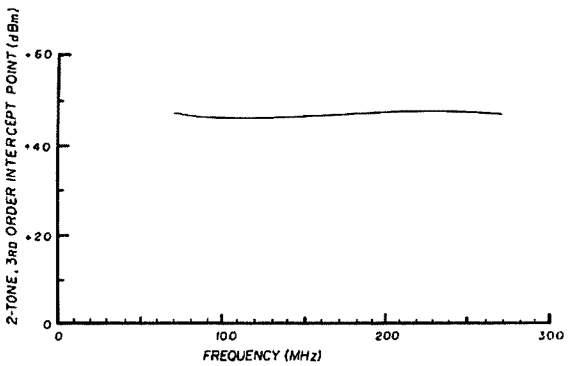


fig. 6. Two-tone, third-order intercept point for the Siliconix VMP4 power fet. V_{dd} is 26V, I_d is 0.4A. Intercept point was referenced to a single tone. Power levels were 100 mW and 1 watt.

taken to know how the numbers were obtained. Some manufacturers may use the PEP output as a reference; others may use average power. A more conservative rating may be obtained by using the single-tone output as a reference, as in fig. 7.

Another interesting feature of the VMP4 *Mospower* fet is that it is not sensitive to load mismatch; there is no need to panic if you disconnect the output cable during testing or tuning. Should a slip occur during a tweaking session, sparks may fly from the metal tool, but when things have calmed down again, the fet will still be ready for action. *Mospower* fets appear to have three fundamental advantages; very easy to match; extremely rugged; and can be operated in parallel without complications.*

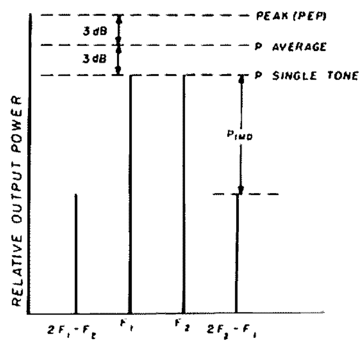
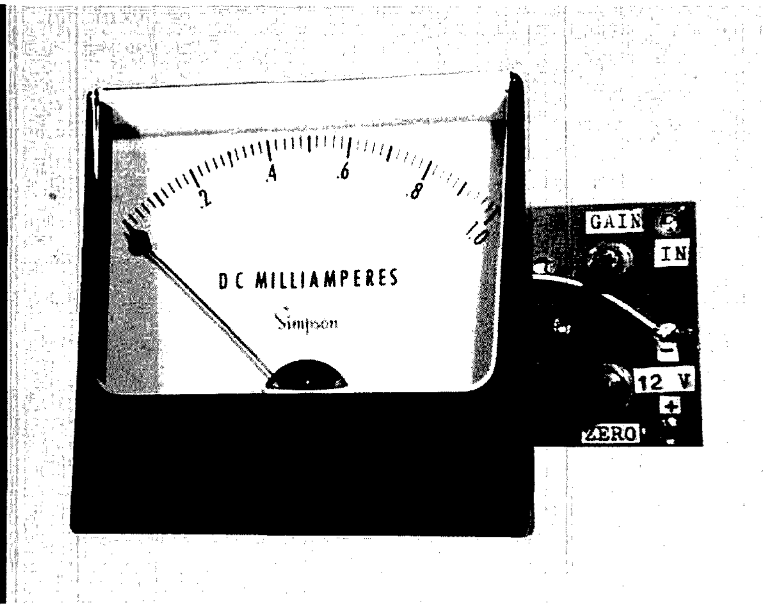


fig. 7. Three common methods of rating IMD performance. The most conservative uses a single-tone output as a reference.

The VMP1 is available for $7.85 each; the VMP4 is priced at $20 each (plus postage and handling charge). California residents please add 6% sales tax. For complete ordering information, write to Ed Oxner, Siliconix, Incorporated, 2201 Laurelwood Road, Santa Clara, California 95054.

ham radio

*Circuits for using *Mospower* fets (VMP1) in single and parallel configurations were given in the September, 1976, issue of *ham radio*, page 10.



electronic meter amplifier

Junction fet in simple circuit turns dc milliammeter into dc microammeter

My **ham radio budget** was seriously threatened recently when circumstances called for use of a 0-100 μA microammeter. Reference to the radio wholesale catalog immediately discouraged the purchase of a new meter. However, a 0-1 mA milliammeter recently acquired at a reasonable price was available from the parts cabinet. The question was, could a 0-1 mA meter, with the aid of a dc amplifier, substitute for a 0-100 μA microammeter? The following description answers the question in the affirmative.

circuit description

A Motorola MPF102 junction fet is used as the active element of a dc amplifier, as illustrated in fig. 1. There are many other types of fets that can be used, including mosfets, depending upon what is available in the junk box. The MPF102 is inexpensive and can provide current gains of over 50 dB in this meter bridge circuit. The maximum full-scale sensitivity of the circuit is approximately 2.0 microamperes. For example, two volts fed to the circuit through a 500k source impedance provides a full-scale meter reading. The 500k gain control is used to set the circuit sensitivity to the desired value.

Parts for the meter amplifier are assembled on a piece of 3/32-inch (2.5mm) thick epoxy fiberglass board. The assembly is attached directly to the rear of the meter by the two meter terminals. I used ¼-inch (6.5mm) brass eyelets equipped with solder lugs and spaced to match the meter terminals. All other terminals are 1/8-inch (13mm) brass eyelets, mounted in 1/8-inch (13mm) holes drilled in the fiberglass board and rolled over with the aid of a center punch and a hammer. Location of the

By Norman J. Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

holes is not critical, but the general layout shown in the photographs can be followed.

Allen-Bradley type-G potentiometers are used for gain and zero-set purposes, primarily because they were available, but standard size potentiometers can be used just as well. The potentiometers are mounted on the right-hand side of the board for accessibility. After the meter and amplifier assembly have been mounted behind a panel, adjustments are made through screwdriver access holes.

adjustment

Adjustment of the meter amplifier is relatively simple. First, with no signal at the input, adjust the zero-set control for zero meter current; then connect the signal to the input and adjust the gain to the desired value. If the source includes a dc offset voltage, it is possible to compensate by properly offsetting the zero-set control.

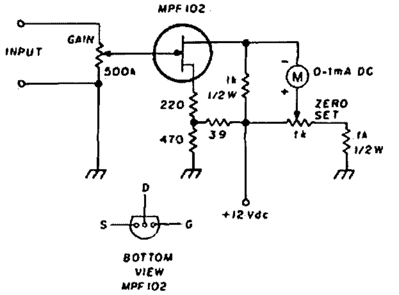
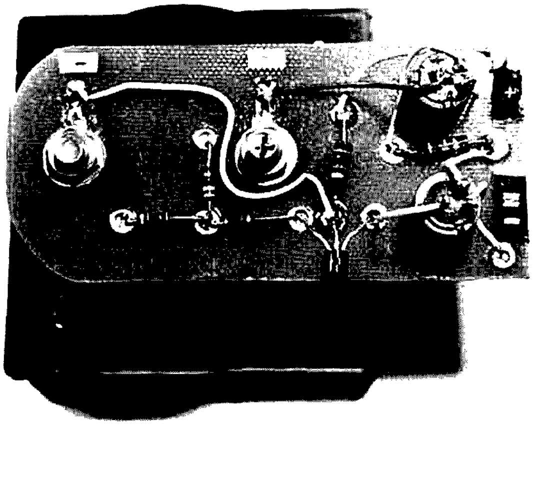
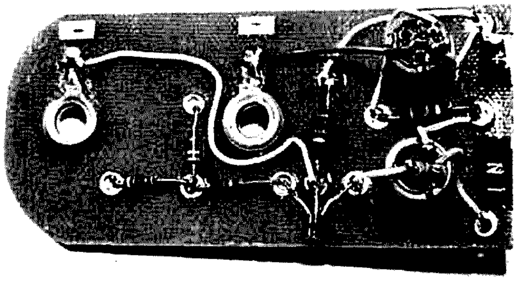


fig. 1. Schematic diagram of the meter amplifier circuit. All resistors are ¼ watt unless otherwise noted.

At my station, the dc meter amplifier is used in conjunction with an rf detector attached to the forward-coupled arm of a 20 dB directional coupler. Since the detector operates in its linear region, the 0-1 mA meter reading is approximately proportional to the current flowing in the main line of the coupler (except at the bottom end of the scale). The meter was calibrated by comparing its reading with an rf power meter borrowed for the purpose. The power meter was connected to the output end of the directional coupler, as shown in fig. 2.



B. Finished circuit board shown mounted to back of meter, held in place by the meter terminals. Neat, simple.



C. Circuit board uses eyelets for component-mounting and convenient tie-points for wiring. Large eyelets are for meter terminals.

Once the scale is calibrated there is no further need for the wattmeter, and antenna power may be read directly from the meter scale with the aid of a calibration chart. In the example given in fig. 2, the full-scale power reading is 1.0 watt. Higher power levels can be monitored with the same setup by using directional couplers having higher coupling values. For example, the full-scale power reading would be 100 watts if a 40 dB directional coupler were used.

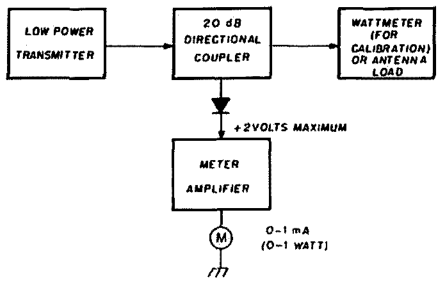


fig. 2. Block diagram showing application of the dc meter amplifier as a sensitive rf wattmeter. In this case 0-1 mA meter reads out 0-1 watt with 20 dB directional coupler (see text).

The dc meter amplifier can also be used as a high impedance dc voltmeter, with a range of 0-1 volt, for example. If the meter amplifier is connected to the avc line in a receiver, it can serve as an S-meter.

references

1. Norman J. Foot, WA9HUV, "Fet Tone Keyer," *QST*, October, 1969, page 103.
2. *Motorola Semiconductor Handbook*, 5th edition, Motorola Semiconductor Corporation, Phoenix, Arizona, page 7-465.

ham radio

ASCII-to-Morse

code translator

Sit back and type near-perfect Morse with this interface between keyboard and transmitter

I suppose every amateur at one time or another has wanted an easy way to send perfect code. The bug, and later the electronic keyer, have to a large degree solved the problem of sending good code. The ultimate way, however, would be to bang out the letters on a typewriter keyboard at any speed while the code is being sent out at some preselected speed. Not only is this nice and easy for you, but the fellow on the other end probably wouldn't complain when he hears your good code. This idea certainly isn't new, but if you've checked lately you know that a commercially produced unit with memory costs about $400, and that's just about out of everyone's price range. This project may just change all that since its unit with memory and keyboard should cost about $120.

code conversion

The initial problem is changing the ASCII code to Morse, but that by no means is the end because Morse characters vary in length from 1 to 6 bits. Since programmable read-only memories (ROMs) are readily available, they're used in this circuit to convert the 6-bit ASCII keyboard output to a special code that includes the Morse letter and its bit length in binary code. The actual code in the memories is not terribly important and is not discussed as most amateurs have no easy way of programming the memories.

Example: The letter R appears at the outputs of the PROMs as (110)00010, where the binary number in parentheses is the bit length (3) of the letter and the right-hand portion is the letter, starting from right to left. Zero is a dot; 1 is a dash. Even though the PROMs can be programmed, I recommend they be purchased

By Robert Morley and Dave Scharon, 2145 East Drive, St. Louis, Missouri 63131

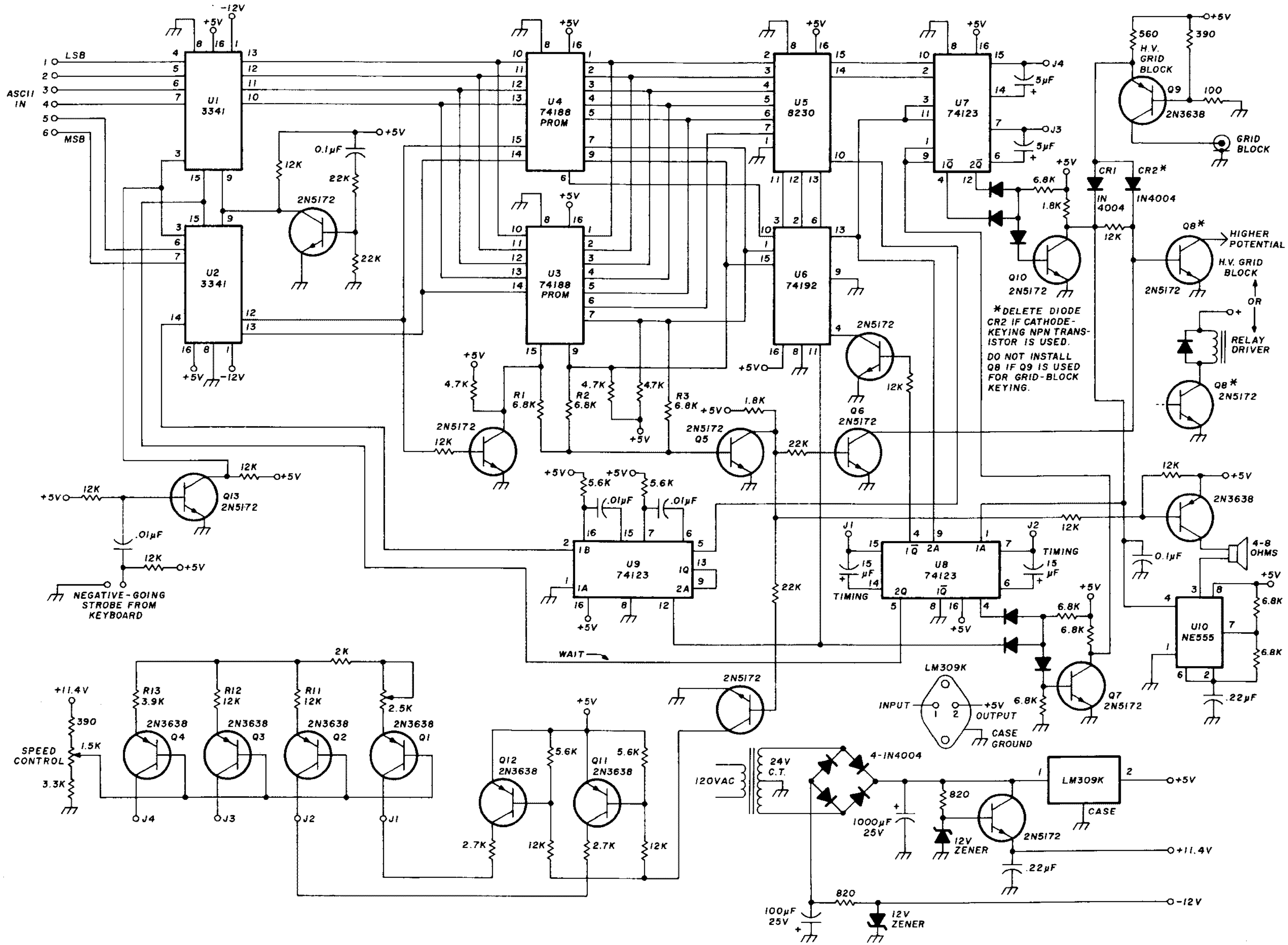


fig. 1. ASCII-to-Morse code translator schematic. U3 is the numeric memory; U4 is alphabetic memory. U1 and U2 are Fairchild 3341 FIFO memories. All resistors are ¼ watt, 10%. Capacitors larger than 1 μF are electrolytics. All diodes not otherwise marked are 1N914 or equivalent.

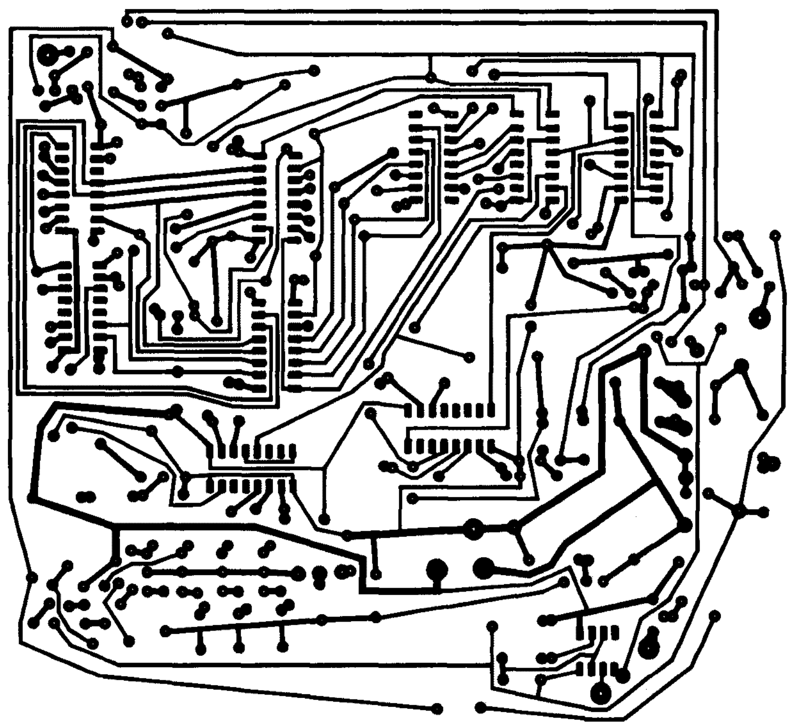


fig. 2. Printed-circuit board layout.

preprogrammed from the supplier listed in the article, because if one mistake is made, that's it. The supplier guarantees correct preprogramming.

The circuit (fig. 1) contains two PROMs. One includes all the alphabet (U4) and the other the numbers and punctuation (U3). Most of the Morse code is represented on a regular ASCII keyboard by its equivalent key except as follows:

key	Morse
<	end of message
>	end of work
=	wait
( )either key	parentheses

This is done because all letters have the 6th bit of ASCII data low and numbers and punctuation high. The 6th bit is used to switch from one PROM to the other. The first 5 bits of both PROMs (U3 and U4) are wired together and connected to U5, an 8230, which is an 8-line to 1-line multiplexer. The last bits address U6, a 74192 up/down counter, which clocks through the bits of Morse letter by selecting the proper lines into U5. The output of U5 determines if a dash or dot is to be fired. The dash or dot is formed by U7, a 74123 dual one-shot. After the dash or dot is formed, the falling edge of the NANDED $\overline{Q}$ outputs fires a space one-shot, U8, whose timing is equivalent to a dot. U8 clocks the counter on the pulse rising edge, which selects the next bit of the letter and triggers the dot or dash on the falling edge. This trigger circle continues until U6 reaches the 000 state. At this time the borrow line of U6 (pin 13) is used to inhibit the dot-dash one-shot, U7, and also fire one-shot U8 to produce a wait between characters equivalent to a dash. This one-shot output is used to bring up the next piece of data in the first-in, first-out (FIFO) memories, U1 and U2.

speed control

It is obviously desirable to have some means to vary the speed of the device, and for simplicity's sake we have chosen one-shots for our dots-dashes and wait timing. Tracking control of the speed would be nearly impossible with variable resistors, as it would take a 4-ganged pot and the control would be nothing close to linear. Current sources, however, make control easy as well as fairly linear. The current sources are controlled by one pot and feed each of one-shots, U7 and U8, at points J1, J2, J3, and J4. These current sources are made up of pnp transistors Q1, Q2, Q3, Q4, in a sort of upside-down emitter follower arrangement. Current control results in good 10:1 speed control.

input buffer

This portion consists of some very interesting circuits, which are the FIFO memories. The input and output are completely asynchronous, which means that data may be entered at any rate independent of the rate at which data is being clocked out by the code generator. As the data is clocked out of the FIFO memory, it is entered

into the two PROMs consisting of U3 and U4. If the keyboard has inverted ASCII, a 7404 ahead of U1 and U2 can be used.

keying circuit

The $\bar{Q}$ outputs of U7, the dot-dash one-shot, are NANDED through a discrete-component NAND gate including Q10. This circuit also drives a high-voltage npn

will normally occur at the end of the last charcter and the end of the H that the space sends, will equal seven dots. This was done by eliminating the space time and one wait time: much greater current is switched in through Q12 and Q11 to the space and wait one-shots while the H is being sent. This action eliminates the time taken up by the three spaces and one wait, which very nearly equals seven dots of time between words.

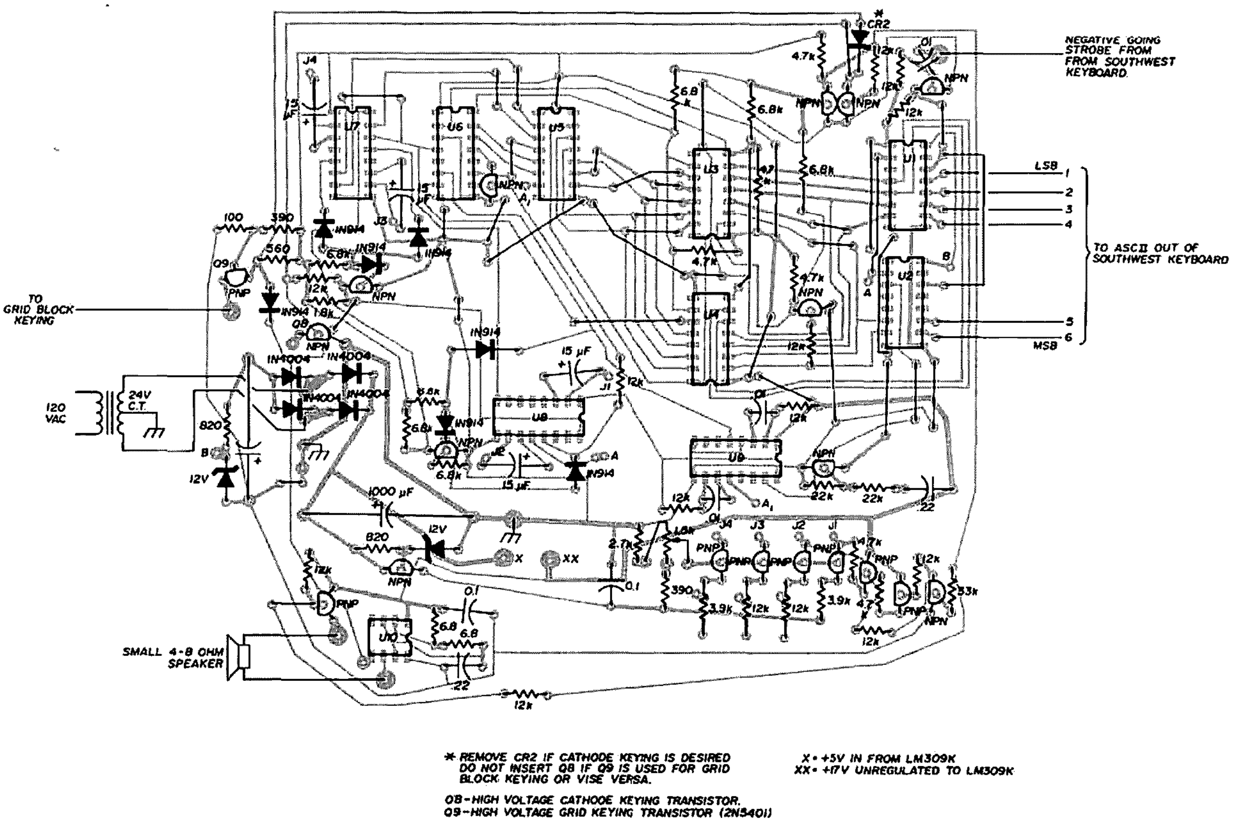


fig. 3. Component side of PC board.

transistor, Q8, for cathode keying and a pnp, Q9, for grid-block keying. The transistors selected here must handle the voltage and current of your transmitter. (This information can be found in the transmitter manual.) Grid-block keying will probably be the most widely used. A good value for the transistor would be 200 volts at 100mA. The NAND output, collector of Q10, also switches a sidetone oscillator made from a NE555, IC10.

space timing

One function that required a trick was the space timing, which is a necessity when the buffer memory is used. This problem required a resistor-transistor OR gate consisting of R1, R2, R3, Q5, Q6, which allows the circuit to function but inhibits the output and sidetone oscillator to simulate the proper space length. When the space bar is hit, a Morse H will be sent. Since good code requires 7 dots between words, something must be done to this H so that its time, plus the two wait dashes that

Any ASCII keyboard will do the job, but be sure it has a bounceless stobe that appears after the data is set up. A good keyboard for this purpose is available.* The circuit is set up for a negative-going strobe into Q13's base. An error key was considered, but when using the buffer chances are you'll be a few characters ahead before you realize an error has been made, and an error signal here would be meaningless. The unit won't make an error itself, so errors should be very rare.

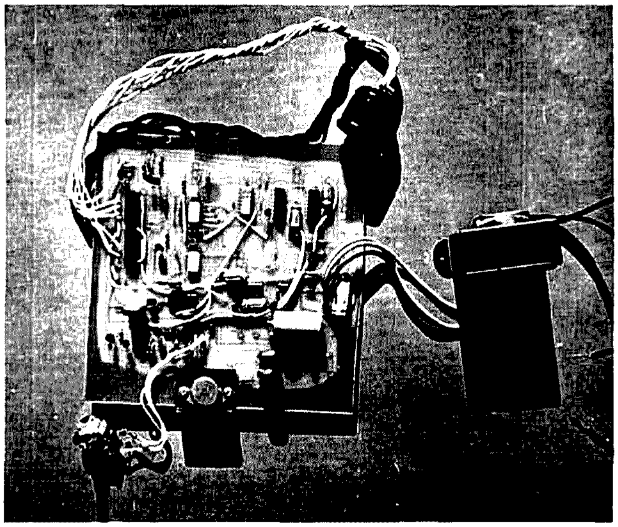
construction

Point-to-point wiring could be used, but a PC board would make assembly easier.† If you prefer to make

*South West Technical Products, 219 West Rhapsody, San Antonio, Texas 78216. (Keyboard kit about $40.00.)

†Circuit boards, 3341 FIFOs, the 8230 and PROMs are available for $60.00 from Scharon Fabricators, 2145 East Drive, St. Louis, Missouri 63131.

your own, **fig. 2** provides a full-size etched board layout; the component side of the board is shown in **fig. 3.** Wiring isn't critical, but care should be taken to ensure proper wiring so none of the integrated circuits will be damaged when the power is turned on. This advice applies particularly to the ROMs and FIFO memories, as they are more expensive than regular TTL integrated circuits.



Printed circuit board for translator chassis showing parts arrangement.

Earlier, the type of possible keying was mentioned. An npn transistor in the Q8 position can be used to key a relay (**fig. 1**). A diode should be placed in reverse across the relay coil to prevent voltage spikes.

The transistors in the circuit aren't critical except for the high-voltage transistors mentioned earlier. The 5-volt power supply is an LM309K. The 3341 FIFOa are MOS integrated circuits, and even though they're internally protected, care should be taken when handling them.

testing and operation

The capacitors used to set the timing on the one-shots for creating the dots, dashes and spaces will never have exact values, therefore resistors R10, R11, R12, R13 may have to be trimmed to get perfect dot-to-dash ratios. Points are available on the board to add these resistors, but in most cases this should not be necessary.

Connect a speaker to the sidetone oscillator and apply power to the unit after the keyboard has been properly attached. (See **fig. 1** for the LSB-MSB keyboard locations.) At this point hit one of the keys on the keyboard and listen for the proper code from the sidetone oscillator. If the proper code isn't heard, check the circuit carefully and if an error is found, try again. Set the speed control as slow as possible, type in a message, then sit back and listen to the code; it should now be ready to put on the air.

ham radio

the ladder network

An analysis of the ladder network using the method of "continued fractions"

If you happen to have an old three-gang tuning capacitor lying idle, the addition of a few resistors and a couple of inexpensive transistors will produce a phase-shift oscillator such as shown in fig. 1: no coils, no fixed capacitors, no variable resistors. This is no precision signal generator but is useful as a handy source of rf signals and has a large tuning range. With 2500-ohm resistors in the ladder network it covers about 250 kHz to 2.5 MHz. With higher resistances the range will be less. Transistors such as the 2N4996 or 2N4274 work well, with 3000 or 4000 ohms between ground and the second collector.

analysis

The main purpose of this article, however, is to show how any ladder network (including the one in the oscillator of fig. 1) can be analyzed in a purely *routine* fashion by the use of a not-too-familiar branch of mathematics called "continued fractions." The analysis applies to ladders composed of any number of elements, each of which can be anything you want. By suitably choosing these elements, the ladder can be made to become not only a phase shifter but also a bandpass, lowpass, or highpass filter; an impedance matching network; an attenuator; or a lumped-constant transmission line.

The input end of the ladder can be a series element to which a constant voltage, E, is applied or a shunt element fed by a constant current, I. The final element can be a shunt element across which the output voltage, e, is taken, or it may be a series element, the current in which i is the output.

For the moment, however, let us concentrate on the case of six elements arranged as shown in fig. 2 where the zs are the impedances of the series elements and the ys are the admittances of the shunt elements. For this network there are two elegantly simple formulas: the output voltage, e, is $\frac{E}{p_6}$ and the input impedance is $\frac{p_6}{q_6}$.

Of course these formulas are useful only when p_6 and q_6 have been expressed in terms of the various impedances and admittances in the ladder. How this is done in a systematic manner requires quite a bit of explanation, but proofs will be omitted so that only what is necessary to operate the mathematical mechanism will be explained. We start by putting all the elements of the ladder in this peculiar-looking form:

$$z_1 + \cfrac{1}{y_2 + \cfrac{1}{z_3 + \cfrac{1}{y_4 + \cfrac{1}{z_5 + \cfrac{1}{y_6}}}}}$$

This expression is called a "simple continued fraction" containing six quantities. Its value is the input impedance of the ladder, although you may not be able to see it offhand. Any such continued fraction can be reduced to the ratio of two expressions involving the zs and the ys, the numerator of this ratio being called p with appropriate subscript, while the denominator is called q. If there had been only one element in the ladder we would have $p_1 = z_1$ and $q_1 = 1$. If there had been two elements we would have $p_2 = z_1y_2 + 1$ and $q_2 = y_2$. But when there are quite a few elements, the reduction to ps and qs would be very laborious were it not for the fact that the theory of simple continued

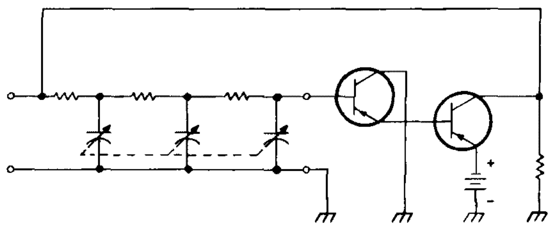


fig. 1. Phase-shift oscillator using ladder-network principle. Capacitor is a three-gang variable; transistors can be types such as the 2N4996 or 2N4274. Oscillator covers about 250 kHz to 2.5 MHz frequency range.

fractions gives us a rule for building up from one p to the next and from one q to the next. The rule is that every new p is equal to the previous p multiplied by the element having the same subscript as the new p, plus the p which is two places behind the new p. Thus, $p_3 = p_2z_3 + p_1$ and $p_4 = p_3y_4 + p_2$ and so on. By the use of this rule a table of ps can be built up without any real thinking. And the same rule applies to the qs starting with the two qs given above. It is not difficult, merely somewhat lengthy, to find the value of p_6 in general terms. But it becomes much easier if we don't make all the elements different. For example if we choose all the zs in fig. 2 to be equal resistances, R, and all the ys to be equal susceptances j/x, where X is the capacitive reac-

By Walter van B. Roberts, K4EA, 6330 Manasota Key Road, Englewood, Florida 33533

tance of one section of the tuning capacitor, then p_6 works out to be

$$\left(1 - 5\frac{R^2}{x^2}\right) + j\left(6\frac{R}{x} - \frac{R^3}{x^3}\right)$$

From this it is evident that when $R/X = \sqrt{6}$ the imaginary part vanishes and the value becomes −29. In other words, the phase has been shifted 180 degrees and the output voltage is 1/29 of the input. Hence if the output voltage is amplified 29 times or more and then fed back to the ladder input, oscillation will occur, as in fig. 1.

the general case of n elements

The foregoing seems sufficient explanation to make the transition to the general case of n elements obvious. **Fig. 3** gives all the information needed to grind out a solution for any ladder with any number of elements of any kind, provided the rule for building up the ps and qs is remembered. But it must be emphasized that the present method does not get around the necessity for writing out lengthy expressions when many different

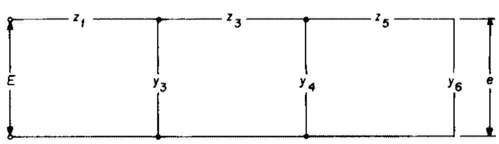


fig. 2. Ladder model using six elements. Network analysis is made in terms of impedances and admittances of the series and shunt elements.

elements are used. What it does do is provide a routine that can be followed without any brainwork other than to be careful not to make simple mistakes.

example

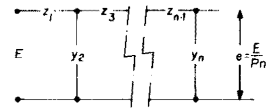
Before attempting to use the material of fig. 3 for anything complicated, however, it's a good idea to get the feel of the mechanism by trying it out on something easy, which can be checked by other methods. For example, in fig. 2 if all the elements are equal resistors, R, it will be found that the output voltage is 1/13 of the input voltage, regardless of the size of R, and that the input impedance is 13/8 times R. Incidentally it may be of interest to note that if the number of resistors is increased much beyond six in the network just discussed, the input impedance approaches a constant value

$$\frac{1 + \sqrt{5}}{2} R$$

Finally, although it should be obvious, the reason that no consideration has been given to ladders with constant voltage applied to a first shunt element, or with constant current fed into a series first element, is that in both these cases the first element becomes functionless. In the first case the input shunt element would merely draw useless extra current from the source of constant voltage, while in the second case the series first element

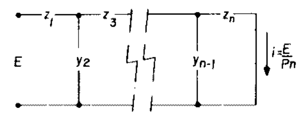
INPUT TO SERIES ELEMENT

INPUT IMPEDANCE = p_n/q_n



The impedance fraction:

$$z_1 + \cfrac{1}{y_2 + \cfrac{1}{z_3 + \cfrac{1}{y_4 + etc.}}}$$



Its ps and qs are:

$p_1 = z_1$

$p_2 = y_2 z_1 + 1$

$p_3 = z_3 y_2 z_1 + z_3 + z_1$

$p_4 = y_4 z_3 y_2 z_1 + y_4 z_3 + y_4 z_1 + y_2 z_1 + 1$

and so on

$q_1 = 1$

$q_2 = y_2$

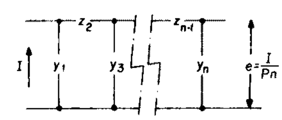
$q_3 = z_3 y_2 + 1$

$q_4 = y_4 z_3 y_2 + y_4 + y_2$

and so on

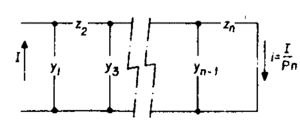
INPUT TO SHUNT ELEMENT

INPUT ADMITTANCE = p_n/q_n



The admittance fraction:

$$y_1 + \cfrac{1}{z_2 + \cfrac{1}{y_3 + \cfrac{1}{z_4 + etc.}}}$$



Its ps and qs are the same as at left with z and y interchanged:

so $p_1 = y_1$

$p_2 = z_2 y_1 + 1$

and so on

fig. 3. Data for solving any ladder network of n elements using the method of "simple continued fractions."

would merely require more voltage in the source of constant current. In neither case would the first element play any part in the performance of the ladder. That would be determined solely by elements beyond the first. Thus fig. 3 covers all actual performance possibilities.

ham radio

repair bench

Joe Carr, K4IPV

resurrecting the old war horse: new hope for the old receiver

If you walk through the fleamarket of any reasonably sized hamfest, it's likely that you will turn up any number of middle age receivers, often at bargain-basement prices. Models such as the venerable Hallicrafters SX-28A (which WA4EPI toted up the side of Bull Run Mountain on Field Day), SX99, SX-100, SX-101, and any of a long series of HQ-series Super Pro receivers by Hammarlund seem to be in evidence. Receivers, especially general-coverage types, manufactured prior to the late 1960s are in poor favor among hams because they lack the gloss of the newer technologies, may be a little troublesome when tuning single sideband (but not necessarily), and don't seem to fit well into the decor of our current transceiver-oriented radio stations. If, however, you are a new novice, a pre-novice wanting W1AW code practice and something to fiddle with, a young adult with a family budget battered by kids, mortgage and a car payment or two, or even an oldtimer with a luxury station, such old war-horse receivers can be a real bargain for use as a standby receiver, the main station receiver, or something to putter around with to learn some shirt sleeve electronics. The kicker is that – they frequently don't work properly.

As I pointed out in my troubleshooting article in the June issue, in most cases receiver troubleshooting is not the terrible chore it is made out to be. To be sure, there are some terrific problems that require a good technician and a lab full of equipment to solve but the vast majority are of a more mundane nature. In this article, I will address those problems that are peculiar to reworking or repairing old, supposedly worn-out, radio receivers which were the big guns of another era. After all, one person's trash may well be another's treasure.

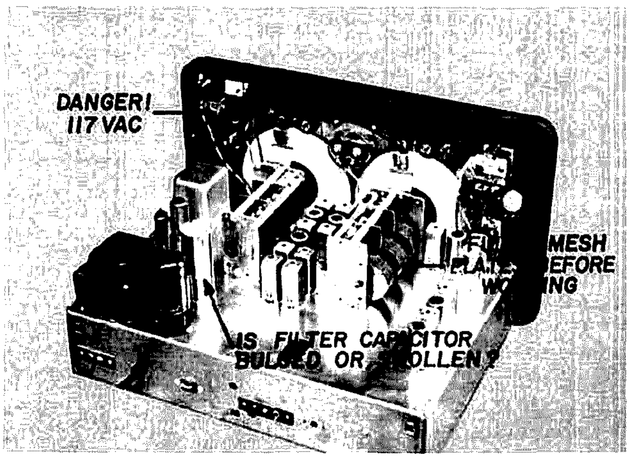
By Joseph J. Carr, K4IPV, 5440 South 8th Road, Arlington, Virginia 22204

Consider my own case. After ten years of apartment living, my wife and I bought a little duplex house and settled into suburban living. While Bonnie fiddled with draperies and furniture placement, Yours Truly was scheming to erect something that had been denied me for all those years: *an antenna.* The next requirement to be met was an old DX-60B belonging to a co-worker and the purchase of a new receiver – until the prices made it apparent that I wasn't going to have a new receiver for a long while. An old friend who never really got into amateur radio was more than willing to part with his old Hammarlund HQ-145X which he had purchased new when Eisenhower was President. It only partially worked, but the price was well below its apparent market value. A decade and a half of improper storage had taken its toll.

The first thing to do when "auditioning" a receiver of this vintage is to make an operational check, preferably before making the purchase. Turn it on and operate the controls. Determine exactly what is, or is not, working properly. Keep in mind that most receivers which have been in storage for any length of time will not be in the best of working order. If the radio works on *any* band, then the set is a good candidate for resurrection. If, on the other hand, it does not work on any band, it is a lesser candidate but that doesn't mean it should be completely ruled out. To be sure, it may require a more extensive evaluation, but the fact that nothing goes "beep" when you tune across the ham bands should reduce the price quite a bit.

Once you have acquired the monster it may have to be repaired. But first, let it run (*not* unattended) for several hours a day over the course of a few days to a week. The heat from the filaments tends to drive out any accumulated moisture, and turning the receiver on for short periods allows the electrolytic capacitors to wake up and start working again. My own HQ-145X had a rather bad audio hum when I first inspected it, but the

fig. 1. The chassis of a reclaimed communications receiver, such as this HQ-145X, may have several points with lethal potentials — unplug the receiver before working on it. Examine the electrolytic capacitors for bulges or a swollen appearance. Make sure the variable capacitor plates are fully meshed before you begin working.



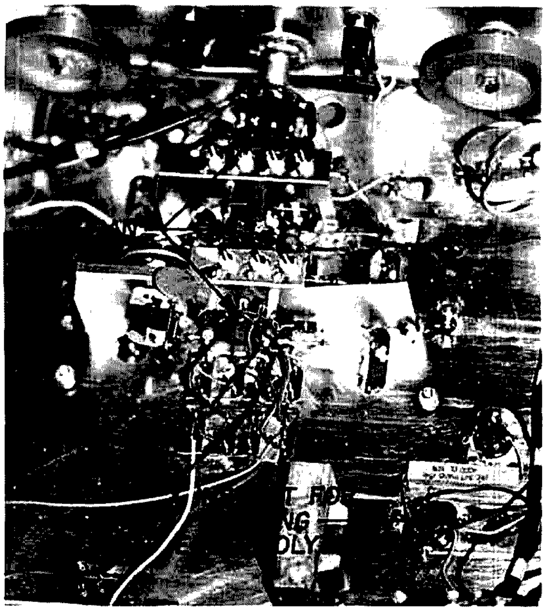


fig. 2. The bandswitch is a main area of difficulty and may need cleaning with a degreaser or a pencil eraser. Examine the bottom of the electrolytic filter capacitor for signs of electrolyte leakage.

hum disappeared in a few days after the electrolytics had reformed. In addition to the audio hum there were a number of other problems:

1. No high band (10 to 30 MHz) and only a scratching sound was heard on these frequencies when the main tuning dial was turned.

2. Occasional oscillation and a rather constant microphonic condition. Oscillation seemed frequency dependent and was at its worst on the special 20 BS band (a modified 10 to 30 MHz band unique to the HQ-145X series).

3. Noisy control potentiometers.

4. Noisy bandswitch.

5. All other rotary switches noisy and/or intermittent.

After the receiver had been burned in for almost a week, I was ready for the next step: cleaning and an additional inspection of the innards. Fig. 1 shows the HQ-145X with the cabinet removed. At this point I feel it's important to caution you that lethal voltages exist on several points inside the receiver. Raw 117 Vac, direct from the power mains, is present on the clock motor in the upper lefthand corner. Unplug the receiver until power is actually needed and discharge the filter capacitors. Also, at this point, be sure to fully mesh the capacitor plates (tune receiver to low end of the dial) so that they will not be damaged as you work. If the capacitor plates are left unmeshed as those in fig. 1, then damage is almost certain.

The first step in cleaning the receiver is to remove the layer of dust that inevitably collects on any electronic chassis. Use either a 1- to 2-inch (2.5 to5cm) paint brush or an air gun, if available. Don't ever use steel wool on an electronic chassis — the small particles can really gum up the works. Although aesthetic considerations and your early training may dictate that you clean every nook and cranny don't be too vigorous in the vicinity of the main tuning capacitor. Dust between the plates can cause trouble and is hard to completely remove.

The next phase of the job is to replace any burned-out lamps, then clean the potentiometers and the rotary switches. The switch wafers and pots can be cleaned with almost any of the spray-can electronic contact cleaners, even the cheapies sold through mailorder and walk-in retail outlets. Be sure to spray each switch wafer separately and try to get spray inside each potentiometer. Immediately after spraying it is wise to vigorously operate the control or switch through its entire range for several seconds. In many cases, though, badly neglected rotary switches, especially those that have been totally unused for years, will have corrosion bad enough (it's black and the spray doesn't cut it) that the spray treatment is insufficient. For switch wafers in that condition I recommend an ordinary pencil eraser applied directly to the contact surface. It is usually best, especially in the vicinity of the bandswitch where moved wires can mean changed alignment, to hold the eraser steady and move the contacts underneath it. Don't forget the portion of the wiper contact surface directly underneath the fixed contacts (see fig. 3). That is, after all, where most of the trouble is.

The initial cleaning just described completely solved problems 3, 4 and 5 on my list and made a dent in problem 1 and now I could hear activity on 20 meters and 15-MHz WWV! The 20BS oscillations remained, as did the scratching on 10 through 30 MHz as the main tuning dial was turned. I also noted that a frequency dependent oscillation had appeared on the 10 to 30 MHz band. This can be a little frightening. If you are cognizant of receiver problems you probably agree that troubleshooting *tunable* oscillations is a lot like trying to

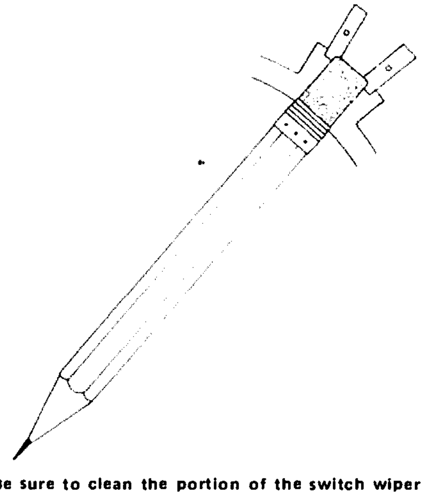


fig. 3. Be sure to clean the portion of the switch wiper contact surface that is underneath the fixed contacts. This can be done by rotating the switch while holding the eraser between two contacts.

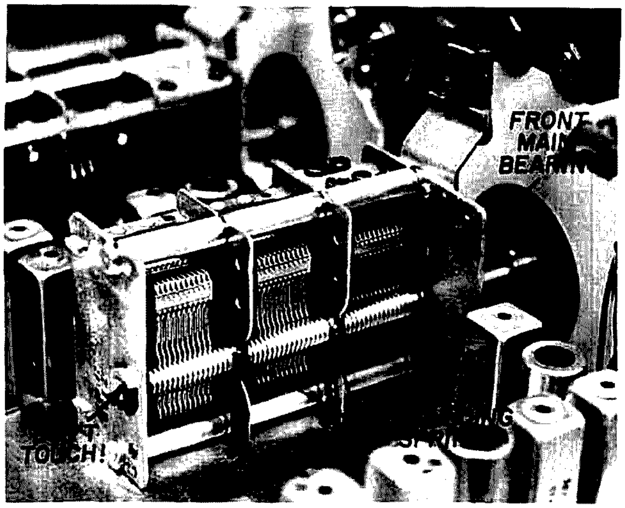


fig. 4. Instability and oscillation results from dirt and corrosion under the rotor ground springs and front main bearing. Clean and relubricate. See text for precautions.

skin an amoeba. I have seen amateur and professional technicians alike (including myself) work for hours, and even days, on such problems. Before looking for open suppressor grids, open screen bypass capacitors, or agc capacitors, you must consider other causes that are peculiar to the neglected receiver.

There are two general causes of oscillation and microphonic conditions on such receivers and they are both spelled *CRUD*. In some cases, it is found that crud on the chassis forms undecoupled feedback paths between stages. Use a Freon-based degreaser or, as an undesirable alternative if you must, that popular concoction consisting of a quart of *Lestoil* with a pinch of ammonia and acetone added, to clean the chassis. Be sure to concentrate on the underside of the chassis and take care not to slop that concoction inside i-f/rf transformers and on to the variable capacitors. The second type of crud is dirt and dried grease in the main bearing and underneath the rotor grounding springs of the main tuning capacitor (see fig. 4). The cleaning should be done with a tiny jeweler's screwdriver or a relay cleaning tool. Carefully burnish the frame and the spring until clean.

The front main bearing race is a circular track filled with ball bearings. Clean this out using a virgin solvent such as Freon TF (I use Miller - Stephenson MS-180). *Do not* use foam contact cleaners or any television type cleaner that leaves a residue. Also, do not trust the labels on some products. Actually test the spray for a residue. Even with a cleaner such as MS-180 use it sparingly and avoid hitting the capacitor plates if possible (it *is*). After the bearing race is clean, re-lubricate it with a white grease such as Lubriplate (available at most electronic supply houses). Use a single dab on the main race and a microdab underneath each grounding spring. Clean the spring over the rear main bearing but otherwise leave that bearing alone. Under no circumstances should you attempt to adjust the tension screw associated with the rear bearing. It is unlikely that the capacitor will ever work properly if you do. Cleaning the main tuning capacitor and the bandspread tuning capacitor (*in* a similar manner) completely cured problem 2 and what was left of problem 1.

Once all of the cleaning is done you may worry about replacing components. Of course, if the radio does not work up to snuff then some troubleshooting may be in order. Chances are good, though, that the cleaning will restore normal operation, assuming that the radio had been retired in good working order.

Examine all elctrolytic capacitors. If any are bulged out or appear swollen, replace them. Also examine the electrolytics for signs of electrolyte leakage. Look for fluid, either a loose, clear stuff or a thick syrup-like type or (more often) a dry powder that will be some color between off-white and dusky brown. Any of these symptoms mark the capacitor for replacement — don't leave a bad electrolytic in the set! If only one section of the *multisection power* supply *filter* capacitor is open do not be tempted by the poor advice, given by some people, to bridge a good capacitor across the bad section. That is only a diagnostic tool and is such poor practice that it ought to be scorned. That open section may short someday (sooner than you might think) and then you can kiss your filter choke and rectifier goodbye.

Examine all of the other capacitors in the set. On paper types look for the wax end plugs being either missing or in poor shape. On the types with a black plastic body look for fluid and cracks in the plastic. Ceramic capacitors may be chipped or cracked but, for the most part, survive well. Replace paper capacitors with a good grade of dipped mylar capacitor such as the

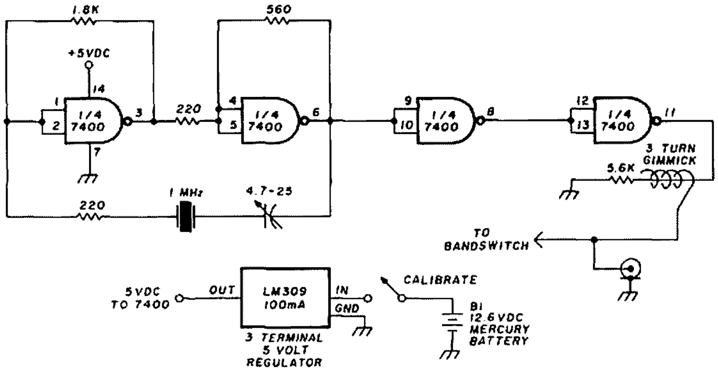


fig. 5. 1-MHz crystal calibrator using TTL logic. The regulator is in the TO-5 can, a 100 mA version of the LM309.

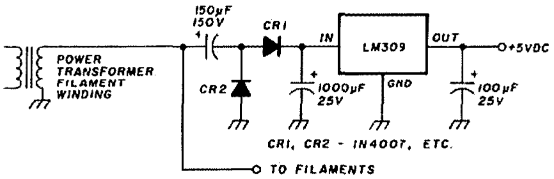


fig. 6. Suggested power supply for the crystal calibrator using the receiver's filament supply.

Sprague Orange Drop. Ceramic and mica types should be replaced with identical units.

Examine all of the carbon composition resistors for signs of overheating, burning, or cracking. Use only a good, new replacement. In this present context let me pass along some advice from my own experience: patronize a quality dealer and buy some good, new, replacement parts. This advice comes from trying bargain, hamfest castoff or surplus parts in one too many circuits! They are good for pittering around, but receiver servicing is really serious work. Incidentally, it won't cost over $10 to $15, even if almost all of the capacitors and resistors in the set are bad, which is unlikely.

You will also want to test the tubes from your reclaimed treasure. Use a tube tester, even if it is in a drug store or supermarket. These are simple emission testers but will quickly spot the gross loss of gain and internal shorts. Replace bad tubes with either new tubes *recently acquired* or those *hamfest types* known to be new military or commercial surplus. I have used a lot of new, surplus JAN tubes over the past two decades and honestly believe that the number of "bad-off-the-shelf" was actually less than those purchased at commercial outlets.

Once your receiver has been repaired and everything works properly, you may want to turn your attention to adding to (or changing) the instrument to fit your own needs. In my case, I activated an old irritation and replaced the three-screw terminal strip used as an antenna connector with a SO-239 coaxial connector. Also added was a binding post connected directly to chassis ground. The receiver which I had acquired lacked a crystal calibrator so one was added. Since I couldn't find a Hammarlund calibrator designed for the HQ-145, one had to be built. You have several alternatives here. You could duplicate the original and plug it into the appropriate socket or hard wire it to the socket. That has the advantage of using the front-panel *calibrator* switch to turn it on. After an inspection of my junk box I chose a 1-MHz TTL calibrator shown in fig. 5. Although mine is battery powered, you could power it from the receiver power supply as shown in fig. 6. The filament drain of the tube in the optional Hammarlund calibrator was about 300 mA and that is more than sufficient reserve to run the TTL integrated circuits.

After all else is done and you know that the radio is going to work, take some mild, soapy, household cleaner and gently clean the front panel. That "new" appearance will give you a psychological boost that is well earned.

ham radio

increased flexibility for the MFJ Enterprises CW filters

Simple circuit changes you can make to the CWF-2 or CWF-3 filters to permit rapid changing of center frequency or bandwidth or both

Perhaps the most popular active audio filters used to improve CW reception are those manufactured by MFJ Enterprises.* The CWF-2 model is an 8-pole filter with selectable fixed bandwidths of 80, 110, and 180 Hz around a center frequency of 750 Hz, as shown in fig. 1. MFJ's other filter, the CWF-3, is a smaller version of the CWF-2, providing a 4-pole filter with 110 and 180 Hz bandwidths.

After using the CWF-2 unit for awhile, I wanted to vary the filter center frequency and bandwidth independently to suit my own preferences. In addition, an article by W6AGW outlined the minimum criteria for the filter bandwidth and center frequency as a function of code speed.[1] Although several articles have described the design and construction of CW filters, I decided to modify my unit rather than build a new one, since the components of the individual stages are already matched to give the same center frequency.

This article describes several modifications that can be made to either the CWF-2 or -3 to permit rapid changing of filter center frequency, bandwidth or both.

basic design

Although W7EIJ[2] and WA1JSM[3] have presented the design steps for such a filter, it is nevertheless worthwhile to briefly summarize their results to better understand what components may be changed. Since each op-amp section (a 2-pole bandpass) is identical to the others, it is only necessary to present the basis for a single stage, as shown in fig. 2, used for Qs less than 10 to minimize ringing. Calculation of component values begins with the selection of C and the choice of center frequency, f_o; stage gain, A_o, at the center frequency; and Q, so that

$$R_3 = \frac{Q}{\pi f_o C} \tag{1}$$

$$R_1 = \frac{R_3}{2A_o} \tag{2}$$

$$R_2 = \frac{R_1 R_3}{4Q^2 R_1 - R_3} \tag{3}$$

For the basic filter section, $f_o = 750\ Hz$, $A_o = 1.32$, and $Q = 4.24$, so that the input impedance is R_1. The bandwidth, Δf, is the frequency difference between the upper and lower −3 dB points, or

$$\Delta f = f_h - f_L \tag{4}$$

where

f_h = frequency of the upper −3 dB point
f_L = frequency of the lower −3 dB point

By Howard M. Berlin, K3NEZ, 2 Colony Boulevard, Apartment 123, Wilmington, Delaware 19802

*MFJ Enterprises, P.O. Box 494, Mississippi State, Mississippi 39762

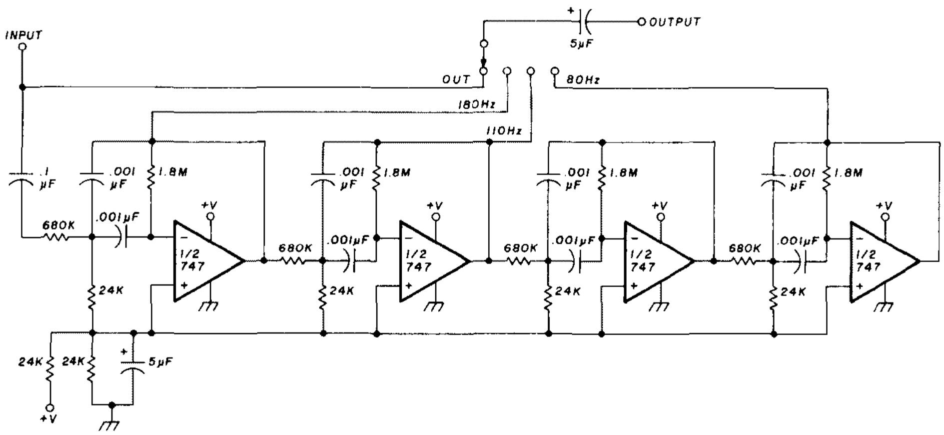


fig. 1. Schematic of the MFJ Enterprises Model CWF-2 CW filter featuring selectable bandwidths of 80, 110, and 180 Hz centered on 750 Hz.

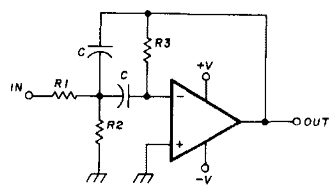


fig. 2. A single-stage, 2-pole bandpass filter section. The calculations of component values for changing filter center frequency and bandwidth are based on this circuit.

The center frequency, in terms of f_h and f_L, is thus

$$f_o = \sqrt{f_h f_L} \tag{5}$$

and

$$Q = \frac{f_o}{\Delta f}$$

$$= \frac{\sqrt{f_h f_L}}{f_h - f_L} \tag{6}$$

The main disadvantage with the circuit of fig. 2, as well as most op-amp circuits, is the need for a dual-polarity power supply. To permit operation with a single-polarity supply, the circuit shown in **fig. 3** is used, with two resistors of equal value for proper biasing. For the MFJ filters **24k** resistors are used, and it's only coincidental that they are numerically equal to R2. In fact, other values could have been used for R.

modification 1

From eqs. **1** through **3**, it can be shown that the center frequency of the single-stage circuit can be changed to a new frequency, $f_o{}'$, without changing A_o or Q by merely changing R_2 to $R_2{}'$, so that

$$f_o{}' = f_o \sqrt{\frac{R_2}{R_2{}'}} \tag{7}$$

Therefore, by using a dual-element potentiometer with series resistors, each combination being equal to R2′, it is then possible to smoothly vary the filter center frequency with two fixed bandwidths using the CWF-3 model. With the components shown in **fig. 4**, the center frequency could be varied from 280 to 1590 Hz with only a 1% variation in either A_o or Q. If the CWF-2 unit is used, a quad-element potentiometer is required for the three selectable bandwidths.

modification 2

If it's desired to smoothly vary the filters bandwidth without varying center frequency, the feedback circuit of **fig. 5**, as suggested by MFJ, can be used with either unit. Only the first stage (180-Hz bandwidth) is held intact while rewiring one of the other op-amp sections. With the component values shown, it was possible to vary the bandwidth from 75 to 150 Hz at the 750-Hz center frequency. Consequently, the filter Q changed from 5 to 10.

modification 3

This final modification is a combination of the previous two and permits the greatest degree of flexibility. You can now select either fixed bandwidths of 180 and 110 Hz or a variable bandwidth with both bandwidths

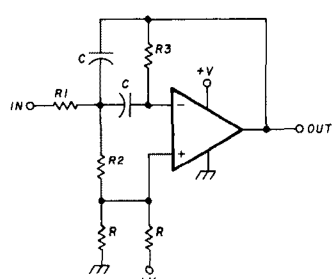


fig. 3. Alternative form of the basic filter stage for use with a single-polarity power supply.

having adjustable center-frequency capability. Using the CWF-2 unit the first two sections are untouched, while one of the remaining op-amp sections is changed. The final circuit is shown in fig. 6.

helpful hints

When soldering on the MFJ printed circuit board, be extremely careful not to overheat the copper laminate, otherwise the copper will separate from the board.

As pointed out by Lancaster in his recently published *Active Filter Cookbook*,[4] ordinary multiple-element potentiometers, particularly the snap-together types,

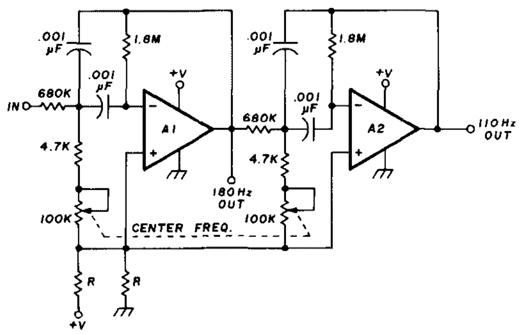


fig. 4. Schematic of the variable center frequency filter with fixed bandwidth. A variation between 280-1590 Hz of center frequency is possible with the two 100-k pots.

have problems. The first is that the resistance behavior is very uncertain at the extremes of pot rotation, since the electrical rotation is somewhat shorter than the mechanical rotation. Also, tracking between the elements should be 5 percent or better.

One inconvenience is linearity; that is, a linear center frequency vs pot rotation change. This is also a problem

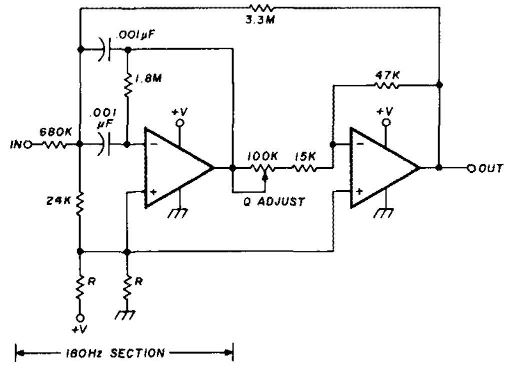


fig. 5. Circuit for varying the MFJ filter bandwidth without varying center frequency. The 180-Hz section is kept intact; the other section is changed as shown to provide bandwidth variation between 75-150 Hz at the 750-Hz center frequency. Filter Q is thus changed from 5 to 10.

with most integrated-circuit keyers.[5] If a linear taper pot is used, the center frequency will change drastically at one end with very little pot rotation. A reverse-log taper should be used for best results; however, multiple-element reverse-log types are both hard to find and expensive.

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ham radio

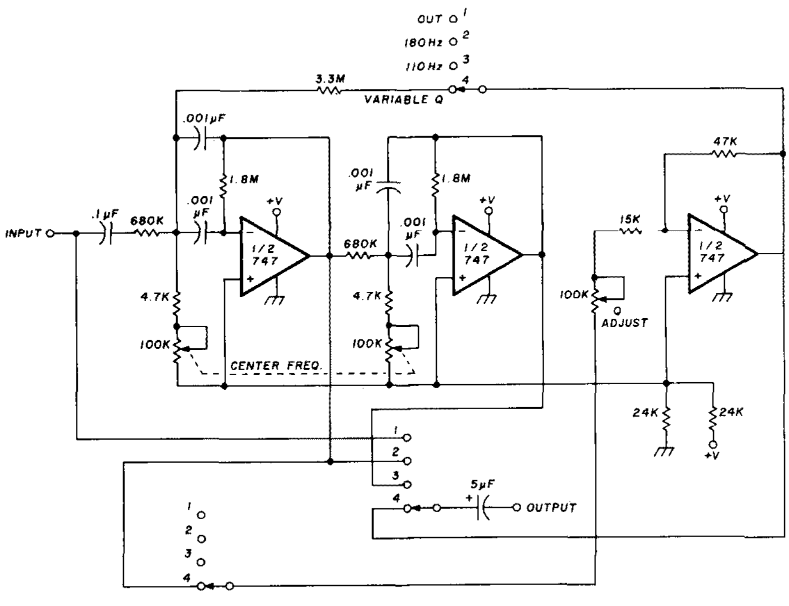

fig. 6. The CWF-2 filter incorporating modifications to permit the greatest degree of flexibility. Circuit features either fixed bandwidths of 180 and 110 Hz or optional variable bandwidth with adjustable center frequency.

RTTY performance

and signal-to-noise ratio

A good case is presented for using low-frequency shift in amateur RTTY systems

Although most RTTY systems on the high-frequency amateur bands use the low-frequency shift of 170 Hz, a few use a shift of 850 Hz. In addition to more mutual interference with CW stations, performance with regard to signal-to-noise ratio (s/n) is not as good with the wider frequency shift. When selective fading is present, the lower-frequency shift will also perform better, as is well known.

The signal-to-noise performance can be shown by using the following formula:

$$\frac{s}{n} = 1.73\frac{D}{f} \cdot \frac{C}{N}\sqrt{\frac{B}{2f}} \qquad (1)$$

where s = signal voltage at output of discriminator and lowpass filter

n = rms noise voltage at output of discriminator and lowpass filter

D = deviation of the carrier from the center to one side (one-half the frequency shift)

f = audio frequency range following the discriminator and lowpass filter

By Leland E. Thompson, K6SR, 14851 Devonshire Avenue, Tustin, California 92680

B = frequency band preceding the limiter (it is assumed that a bandpass filter is used)

C = carrier voltage in frequency-band B before limiter

N = rms noise voltage in band B

Assuming a speed of 60 words per minute, the following values can be placed in eq. 1:

$B = 250\ Hz$

$f = 28\ Hz$

$D = 85\ Hz$ (total frequency shift is $170\ Hz$)

Then $s/n = 31.2\ dB$ (when C/N is $10\ dB$).

Threshold in wideband fm systems is considered to exist when C/N is 10 dB. In narrowband systems such as those considered here, the threshold is not as sharply defined, and good performance can be obtained with this ratio several dB lower. The s/n of 31.2 dB is 11 or 12 dB more than necessary for satisfactory performance.

The following values can be used in eq. 1 for wide-shift systems:

$B = 850\ Hz$

$f = 28\ Hz$

$D = 425\ Hz$ ($850\ Hz$ total shift)

The $s/n = 50.2\ dB$ (when C/N is $10\ dB$). This value of s/n is far higher than necessary, and it decreases fast as C/N drops below 10 dB. Since N is 5.3 dB higher because of the wider frequency band, C must also be 5 dB higher. With a fading signal, the wideband system will start producing errors before the narrow system will.

Although there may be a slight advantage in reducing the shift below 170 Hz, band B would have to be reduced by using a narrower filter, and stability would then become more of a problem.

Equal performance should be obtained with the phase-locked RTTY terminal unit described by Webb.[1,2] In this case, band B is determined by the phase-locked-loop bandwidth and f is determined by the low-pass filter, as in the previous system.

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ham radio

MICROPROCESSORS

microcomputer interrupts

This month's column is the first of several that will focus upon the concept of an *interrupt*. When used in the context of a computer, an interrupt can be defined as the suspension of normal program execution in order to handle a sudden request for *service*, i.e., assistance, by the computer. At the completion of interrupt service, the computer resumes the interrupted program from the point where it was interrupted.[1] This specific use of interrupt is consistent with the general meaning of the term: to stop a process in such a way that it can be resumed.

A given computer will typically communicate with a variety of external I/O "devices." If a minicomputer is used, it may communicate with a teletype or alphanumeric keyboard, a CRT display, a printer, a floppy disk, and perhaps one or more laboratory instruments. If it is a microcomputer, it may communicate with smaller devices — (motors, solid-state relays, push-button switches, display lights) within a larger machine or instrument. When used as a replacement for discrete logic devices in a complex digital circuit, a microcomputer may communicate with other TTL integrated circuit chips such as latches, flip-flops, and three-state buffers.

When communicating with external I/O devices,[2] microcomputers can operate in two general modes, *polled* and *interrupt*. Polling is the periodic interrogation of each I/O device that shares a communications link to a microcomputer to determine whether it requires servicing. A microcomputer sends a poll that has the effect of asking the selected device, "Do you have anything to transmit?", "Are you ready to receive data?", and similar questions. When a microcomputer services a polled device, it simply exchanges digital information with the device in a manner that is prescribed by software in a subprogram or subroutine called a *software driver*.

By Peter R. Rony, Jonathan A. Titus, and David G. Larsen, WB4HYJ

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering, are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus is President of Tychon Inc., Blacksburg, Virginia.

In a polled operation, the microcomputer sequences through the devices tied to the microcomputer, looking for individual devices that need servicing. When it finds a device that requires service, sequencing stops and a software driver services the device. Once it is finished, the microcomputer continues checking the devices. Polled operation is most useful with relatively slow devices that do not require frequent service, do not require attention from the microcomputer for excessive periods of time, or can wait to be serviced. Advantage is taken of the difference in speed of operations in the microcomputer and operations in the I/O device. Most common I/O devices are much slower than microcomputers. For example, in 100 ms (teletypewriter response time) an 8080A-based microcomputer can execute approximately 20,000 instructions when operated at a clock rate of 2 MHz. Although a microcomputer may give you the impression that it is doing several things simultaneously, this is only an illusion since it can manipulate data much faster than most I/O devices can respond to changes in data. *A single computer can perform only one task at a time.*

In interrupt operation, the microcomputer juggles the demands of the external I/O devices. There is a distinction between slow devices that require infrequent servicing and high-speed devices that demand the attention of the microcomputer for most of the time. The most appropriate description for interrupt operated systems is that they are *asynchronous*, i.e. they lack a common synchronizing signal and therefore give rise to generally unexpected or unpredictable program execution within the microcomputer. An *asynchronous device* is a device in which the speed of operation is not related to any frequency in the system to which it is connected.[3] The use of asynchronous devices is the rule rather than the exception.

There can exist *priority* in interrupt operation. All I/O devices can be ordered in importance so that some devices take precedence over others. In contrast, there is usually no priority in polled operation. Once a device is serviced, it waits its turn until all other devices are sequenced and, if necessary, also serviced. The time between the interrupt request by a device and the first instruction byte of the software that services it is known as the *interrupt response time*. For a high-speed device that has high priority, the response time can be very short, less than a millisecond. For a low-speed device that has low priority, the response time is variable, since it depends upon the demands placed upon the microcomputer by all higher priority devices.

Three commonly used microcomputer interrupt techniques are the *single-line interrupt*, the *multilevel interrupt*, and the *vectored interrupt* (fig. 1). In the

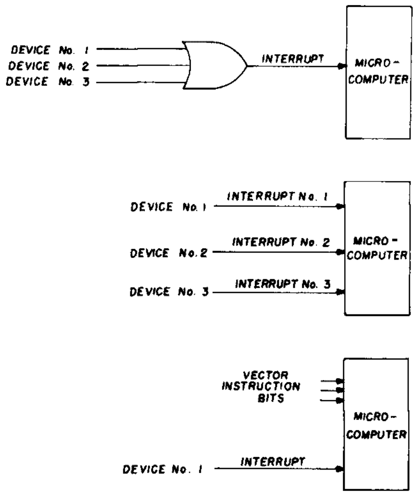

fig. 1. Schematic diagram of three different interrupt techniques, the polled interupt (top), multilevel interrupt (middle), and vectored interrupt (bottom).

single-line interrupt technique, multiple devices must be OR connected to a single interrupt line to the microcomputer. Once an interrupt signal is received, all of the interrupt devices are polled to determine which one caused the interrupt. It is possible to assign software priorities to the various interrupting devices, so that the first device polled that needs service is the one that receives the attention of the microcomputer. A common term used for the part of a program that polls interrupt devices is a *flag checking routine.* We shall discuss the concept of a flag in a subsequent column. At the moment, consider a flag to be a single-bit memory that indicates when an operation has been completed or when a condition has been attained.

In the multilevel interrupt technique, there are several interrupt lines to the microcomputer, each line being tied to a separate I/O device flag. The microcomputer does not need to poll the devices to determine which one caused the interrupt. This is done internally within the microprocessor chip. Depending upon the nature of the microprocessor chip, this can be a very fast interrupt technique, but it is somewhat difficult to expand.

vectored interrupt

A vectored interrupt causes a direct branch by the microcomputer to that part of the program that services the interrupt. This interrupt technique requires external integrated-circuit chips to supply the memory address of the *interrupt service routine* as well as to set the priority. With the 8080A microprocessor chip, eight different service routine addresses can be readily specified, although one of these addresses coincides with the reset address for the microprocessor, location zero. If you are interested in vectored interrupts, we encourage you to consider the Intel 8259 programmable interrupt controller, which became available commercially in July, 1976.

The use of interrupts should be considered very carefully. More complicated software is invariably required. For example, you will generally have to save the status of the microprocessor chip at the time that the interrupt occured. This means placing the contents of the accumulator, the flags, and the registers into a specified region of memory from which they can be retrieved at a later time, after the interrupting device has been serviced. Pay attention to priorities. Make certain that devices that require high priority and need immediate servicing are given the highest priority. Other devices, such as teletypes, should be low priority. Also, if you attempt to do too much with an interrupt system, you might find that your microcomputer becomes "interrupt bound," which means that the microcomputer is only working on the main task, which it should be doing while only infrequently servicing interrupt requests.

To end this column, we would like to provide one example of an interrupt system. Assume that your microcomputer is performing mathematical computations on 7-bit ASCII numbers that are entered via a UAR/T chip[4] that is connected to a Teletype operated at 110 Baud, or ten ASCII numbers per second. The exchange of data between the microcomputer and the UAR/T can be performed in 20 to 30 microseconds, which leaves 99.97 ms left for the microcomputer to do other things. With the Intel floating point package, for example, each floating-point multiplication or division can be performed in 2 to 5 ms with an 8080A-based microcomputer operating at 2 MHz. Sixteen-bit binary multiplications and divisions can be performed even faster. Therefore, it is appropriate for you to consider that the main task of the microcomputer is to perform such computations, and that 0.05 to 0.10 percent of the time the microcomputer can devote its attention to servicing the interrupting teletype. The less attractive alternatives are for the microcomputer to either poll the UAR/T or else to wait for a change of state of the UAR/T data ready or transmitter buffer empty flags.

Interrupts are also effective for use with devices that provide data to a microcomputer but which have no buffer of their own to store it. Existing data must be removed from the device and stored in the microcomputer quickly before a new data word can be generated by the device. One example of such a device is an analog-to-digital converter (ADC) in which the conversions are clocked by an external clock at repetitive time intervals.

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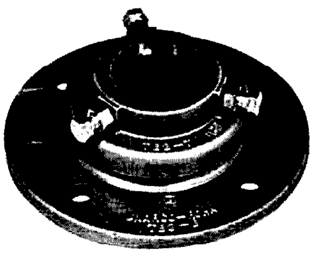
ham radio

NEW products

antenna mount thrust bearing



Unarco-Rohn has just introduced a new thrust bearing (TB-3) for mounting antennas. Manufactured from heat-treated cast aluminum for extra strength, the TB-3 incorporates 30 stainless-steel ball bearings in a race that is protected from the elements and permits free movement at all times. Three lock-nuts fasten the antenna mast securely to the thrust bearing, relieving the weight of the antenna on the rotor and allowing an exceptionally free turning movement.

For additional information, write Rohn Products Advertising Agency, Box 2000, Peoria, Illinois 61601 or use *check-off* on page 142.

low- and medium frequency radio scrapbook

Ken Cornell's new scrapbook fills a huge gap with good, basic, hard-core information. The gap? That range of radio frequencies extending from the bottom of the broadcast band down to about 10 kHz – nearly audio!

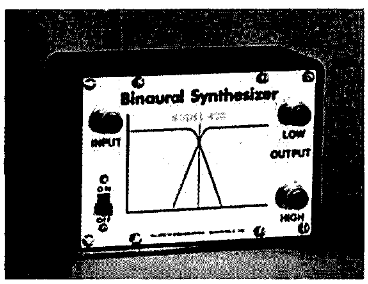
There's another gap, too: a gap in knowledge and information about regulations, communications capabilities, equipment, and "personalities" of those bands below the broadcast.

To all but a mere handful of dedicated experimenters such as Ken and his fellow "LFers," the low and medium frequencies today are nearly as much of a "no-man's band" as they were 50 years ago. *The Low- and Medium-Frequency Scrapbook* not only bridges that gap, but tends to fill it with circuits for practical receivers, transmitters, converters, antennas, and neat little doodads that you will find indispensable for your experiments.

Like any good scrapbook, this one presents the material in such a way that you can find what you need in chapters about communications bands where no license is needed, about coils and coil-winding techniques, and many other comparable "goodies." Tables, charts, nomographs and schematic diagrams are liberally sprinkled throughout its pages.

Even better, the pages are loose-leaf and already punched for three-ring binders so that your scrapbook can be a growing, personal and up-to-date source reference with pages of your own experiments and results added where appropriate. You need this book! Soft cover, 114 pages, $6.95. With custom loose-leaf binder, $8.95. Order from Ham Radio Books, Greenville, New Hampshire 03048.

binaural synthesizer for audio signals



Hildreth Engineering Company announces its new Binaural Synthesizer, a device for producing a stereo audio output from a monaural audio input. The synthesizer may be plugged into the headphone or speaker jack of a receiver or other monaural audio signal source and provides outputs for stereo headphones, power amplifiers, or speakers.

The model 400 Binaural Synthesizer is available as a complete unit (less batteries) for $29.95, as a partially assembled unit including an assembled and tested PC board for $17.95, or as a kit which includes a predrilled, plated, glass-epoxy PC board with instructions for $6.95. All prices postpaid.

The model 400 provides a sharp 24 dB per octave channel separation, four poles in each channel, for a good stereo effect. The unit employs 741 integrated-circuit op amps rather than multiple units for simple, easy-to-find attachment points. The PC board contains extra traces for resistor trimming, if desired. The cross-over frequency is 750 ±50 hertz, input impedance is 2000 ohms, and the unit may be powered from any supply that will provide negative or positive voltage between 4 and 15 volts dc. In operation, the synthesizer typically draws 5 to 10 milliamperes.

Additional information may be obtained by writing to Hildreth Engineering Company, Post Office Box 3, Sunnyvale, California 94088, or by using the *check-off* on page 142.

utility case for test instruments

Continental Specialties Corporation (New Products, July, 1976, *ham radio*) makes available to the custom designer a matching blank utility case that will accommodate CSC's line of test instruments. The DM-C *Design Your Own Mate* case is molded from flame-retardant ABS plastic which can be easily drilled, routed, sawed, filed, or reamed. The blank box incorporates a bottom plate with mounting screws and features a sloping front panel which allows easy custom design and provides a modern, utilitarian appearance. The DM-C is 6.75 inches (17.2cm) long, 7.5 inches (19cm) wide and 3.25 inches (8.3cm) high with a nominal wall thick-

ness of 0.085 inch (2mm) and sells for only $5.50. Write Continental Specialties Corporation, 44 Kendall Street, Post Office Box 1942, New Haven, Connecticut 06509, or Box 7809, San Francisco, California 94110, or use *check-off* on page 142.

range of HP counter extended to vhf and uhf

Designed with the needs of the communications industry in mind, this new Hewlett-Packard Model 5305B Frequency Counter plug-on covers frequency bands from 50 Hz to 1300 MHz. In addition to all mobile communications bands, the counter's range includes TACAN/DME and ATC radar transponders as well as am and fm broadcast bands and vhf and uhf television bands.

The 5305B operates with the 8-digit 5300B mainframe. Sensitivity is 20 millivolts rms in both channels over the full bandwidth. A high resolution mode for tone measurements to 10 kHz improves resolution by 1000. Automatic gain control is included on both channels, plus a manual attenuation control on the high frequency channel.

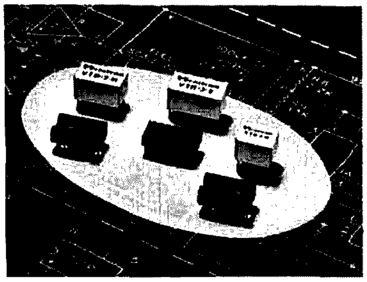
A first for a counter is a probe power output to drive an accessory preamplifier. The Hewlett-Packard Model 10855A Preamp boosts sensitivity of the counter by 22 dB (x10). Adding an antenna and tunable filter to the preamp lets the user receive "on-the-air" signals for carrier measurements.

U.S. price of the Hewlett-Packard Model 5305B is $900, the Model 5300B mainframe is $460, and the Model 10855A Preamp is $225. Delivery for any of these units is 30 days.

For additional information contact Inquiries Manager, Hewlett-Packard Company, 1501 Page Mill Road, Palo Alto, California 94304; telephone (415) 493-1501, or use *check-off* on page 142.

More Details? CHECK—OFF Page 142

miniature ceramic filter for communications receivers



Vernitron Piezoelectric Division recently introduced a new series of low-cost miniature VTD ceramic ladder filters designed to meet the needs of modern double conversion systems.

The small size of the VTD filters and their high stopband rejection, make them particularly suitable for hand-held transceivers, CB equipment, police scanners and commercial two-way radios.

These compact, highly sensitive ceramic filters are fully compatible with transistorized amplifier circuitry, as well as with ICs in am and fm sets. They plug into 14- or 16-pin DIP sockets or mount directly on PC boards, and provide maximum design flexibility.

Vernitron has designed the new VTD Series of ceramic ladder filters in three model groups. VTD-1 models offer 40 dB bandwidth and stopband of 25 to 27 dB; VTD-2 models offer 50 dB bandwidth and stopband of 30 to 40 dB; and VTD-3 models offer 60 dB bandwidth and stopband of 40 to 45 dB. The VTD-1 and VTD-2 filters plug into a 14-pin dip socket and the VTD-3 filter plugs into a 16-pin dip socket.

For more information and Data Sheet 940 write Mark Rickman, Vernitron Piezoelectric Division, 232 Forbes Road, Bedford, Ohio 44146; telephone (216) 232-8600, or use *check-off* on page 142.

Sinclair offers free brochure for amateurs

Sinclair Radio Laboratories — the company known for quality and innovation in the radio antenna systems equipment market for over 25 years —

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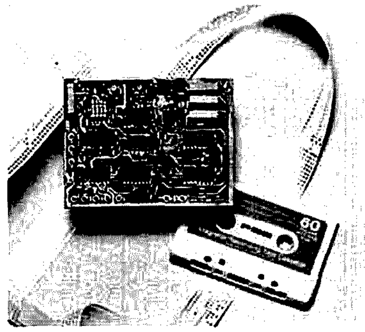
announces that radio amateurs are now able to obtain a *free* "Equipment for Amateur Radio Service" brochure.

Sinclair is the originator of the *Excalibe*r mobile antenna and the tunable *Mirage* antenna, and is responsible for developing the two-meter 100 dB hybrid ring duplexer, the Q-circuit© duplexer and the popular Q-filter.©

Now amateur radio organizations and individuals may obtain all Sinclair products at generous professional discounts.

To obtain the new brochure with price list, discount structure and order form, write: Amateur Brochure, Sinclair Radio Laboratories, Inc., 675 Ensminger Road, Tonawanda, New York 14150; or use *check-off* on page 142.

low-cost audio cassette/tty/crt adapter for microprocessors



Electronic Product Associates announces the availability of a new, low-cost audio cassette/TTY/CRT adapter which allows any serial TTL or MOS output to simultaneously interface a low-cost audio cassette player via frequency shift keying (Byte Standard) up to 300 Baud and to a standard RS232 CRT and a 20 mA current loop TTY. The adapter also simultaneously decodes Byte Standard fsk data from low-cost audio cassette players and from 20 mA current loop TTY and RS232 CRT. Audio cassette information is decoded by a proprietary phase-locked-loop system developed by EPA which is said to be the most reliable method available for transferring digital data to and from low-cost audio cassette players. The Model TCC3 is 4½x3¼ inches

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(11.5x8cm) and mounts piggy-back on the EPA Micro-68 development computer. The TCC3 price is $129.00 in singles, completely assembled and tested. Delivery is from stock.

For additional information, write Electronic Product Associates, Inc., 1157 Vega Street, San Diego, California 92110; telephone (714) 276-8911 or use *check-off* on page 142.

IC Zener has one-ohm dynamic impedance

National Semiconductor Corporation recently announced a new reference diode with a dynamic impedance two orders of magnitude less than that of discrete Zener diodes. The LM129 linear IC, 6.9-volt, reference diode operates over a 0.5 to 15mA current range, allowing it to replace a wide variety of discrete devices and improve circuit performance. An important feature of the LM129 is that all operating characteristics of the reference are essentially independent of operating current.

The heart of the linear integrated circuit is a new sub-surface-breakdown Zener that yields a very low noise and highly stable breakdown. Long-term stability is typically 20 ppm while noise is guaranteed to be less than 20 μV. Active circuitry around the Zener buffers external current changes to give a one-ohm dynamic impedance.

According to Robert C. Dobkin, inventor of the LM129, this new IC Zener will greatly simplify bias circuitry needed to make a reference. Normally, a precision current source is needed to bias reference Zeners; however, with the LM129, only a single resistor is needed.

The new reference is available in selected temperature coefficients from 0.001% to 0.01% per degree C for use in a zero to 70 degree C, or −55 to 125 degree C temperature range. The new IC is packaged in either a TO-46 hermetic transistor package or a plastic TO-92 package.

Pricing for the LM129AH (0.001% per degree C, −55 to 125 degree C) reference is $15.00 in quantities of one hundred, and the LM329DZ (0.01% per degree C, zero to 70 degree C) reference is $0.75. Delivery of both IC reference

diodes is from stock. For additional information, including typical circuit applications and wiring diagrams, write Roy Twitty, National Semiconductor Corporation, 2900 Semiconductor Drive, Santa Clara, California 95061; telephone (408) 737-5287, or use *check-off* on page 142.

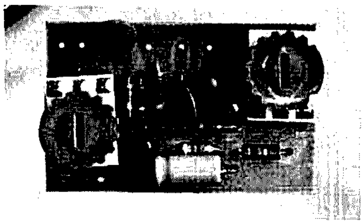
B&K-precision issues new 40-page catalog

A new 40-page catalog describing B&K-Precision test instruments has just been released.

Among the new "cost effective" instruments in this catalog are a 30 MHz scope, a 15 MHz scope and a 5 MHz scope. A new low-cost 3-1/2 digit multimeter is also being introduced for the first time. Other devices described in the catalog include frequency counters, signal generators, semiconductor and transistor testers and numerous other test instruments.

The catalog is available without charge by writing directly to Mr. Paul Mangione, B&K-Precision, Dynascan Corporation, 6460 W. Cortland, Chicago, Illinois 60635; telephone (312) 889-9087, or use *check-off* on page 142.

Touch-tone pad



Touch-toners, if you've been looking for a neat way to interface a touch-tone pad with your two-meter rig, Trevor Industries may have just the device you need. The "Trevorface" has an adjustable audio output level and, each time a tone button is depressed, automatically keys your transmitter for any desired period of time up to several seconds. The printed-circuit board with components is small enough to find a home inside the enclosure of any but the smallest hand-held rig and measures only 3/8x1¼x2 inches (10x32x51mm). The design was developed nearly four years ago for the TR-22 by W2EUP who

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made "GLB" famous, and a large number of these tiny units have been used successfully by amateurs in the Buffalo, New York area for several years. Five simple attachments to B+, ground, PTT line, and audio make hook-up to almost any pad simple and fast, and the unit even includes a 10-volt Zener diode to eliminate alternator "whine." Priced at only $7.95 plus $.45 for postage and handling (New York State residents please add 7% sales tax), the "Trevorface" fills a definite need. Want one? Write or call Gary Ketch at Trevor Industries, Inc., Box 102, Getzville, New York 14068; telephone (716) 834-1639, or use *check-off* on page 142.

ham radio operating guide

Most amateurs like to be known as good operators, and The American Radio Relay League has just introduced a new book that will help the newcomer and brush up the old timer's operating habits. *The ARRL Ham Radio Operating Guide* is an easy-to-read manual for introducing the reader to the many and varied operating practices that exist throughout amateur radio.

The book is written by experts in each of the fields covered and answers questions faced by beginners in any phase of ham radio: What's the best Novice daytime DX band? What crystal frequencies should you order for your new fm transceiver? In the 75-meter phone band, where are Australian stations most likely to be found? When will an amateur satellite be in range of your station? What is the WAJA award? How long should you make each transmission when trying to call another station by meteor scatter?

The guide contains ten chapters: Getting Started, Message Handling, Contests, DX, Awards, Repeaters, Flea Power, Communicating Visually, VHF/UHF, Searching for New Horizons, Oscar.

The manual contains 128 pages and measures 8-1/4 by 11 inches. Price: $4.00 in the United States and its possessions and $4.50 elsewhere. Order your copy today from Ham Radio Books, Greenville, New Hampshire 03048.

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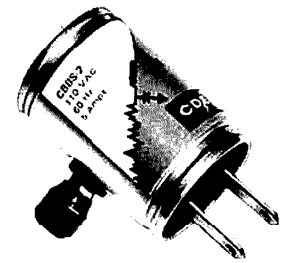
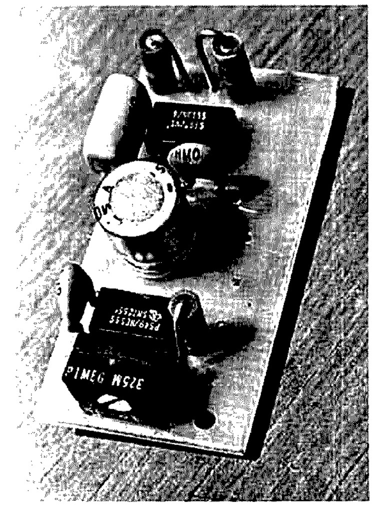
appliance noise filter



Cornell-Dubilier has announced a new and efficient filter, the CBBS-1, for electrical interference produced by hot combs, blenders, electric shavers and similar appliances. Simply plug the Cornell-Dubilier CBBS-1 into the wall socket and then plug the noisy appliance into the CBBS-1. All annoying electrical noise which hampers radio and TV reception is removed. The unit is lightweight, highly reliable, (handles 5 amperes), and is built to resist the effects of heavy household usage.

For additional information, contact William Carlson, Cornell-Dubilier, 150 Avenue L, Newark, New Jersey 07101, or use *check-off* on page 142.

new repeater time-out timer



A new integrated circuit timer for repeater time-out use has been announced by Timekit.® All circuitry of Timekit's Model TG970 time-out timer is on a printed-circuit board measuring only 1x2 inches (25.4x50.8mm) that should fit inside most transceiver enclosures. External

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connections to the TG970 are 13 volts dc, PTT, ground and output.

When the microphone button is pressed (or when the COR is energized) a timing cycle continuously variable from 10 seconds to 30 minutes is activated. At the conclusion of the cycle, that is, at time-out, the TG970 produces a 13-volt, 2 Hz pulse train which can be used to flash an LED, activate a beeper or excite a Timekit Model TG3 oscillator designed for attachment to the TG970 board. The oscillator will pulse an external speaker at 2 Hz with a 1000 Hz note. Timer reset is immediate with no false triggering, and is immune to rf.

The TG970 is priced at $7.95 wired and tested, or $6.50 in kit form. The TG3 is priced at $4.50 wired and tested, or $4.00 in a kit.

For additional information, write Timekit, 23715 Mercantile Road, Cleveland, Ohio 44122; telephone (216) 464-3820, or use *check-off* on page 142.

solderless coaxial connector

The Bunker Ramo RF Division has introduced a PL-259-type Amphenol connector that provides instant and simple termination of RG-58A/U coaxial cable without solder, special tools, or adapters. Designated Amphenol 85-58FCP (for field crimp plug), the new, reusable connector has application in both fixed and mobile stations wherever coaxial cable termination must be made.

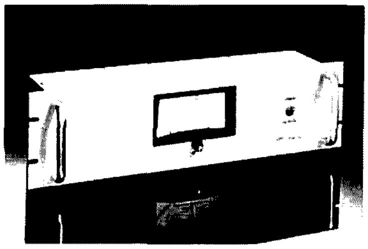
To complete a termination, the user simply strips the coaxial cable and pushed the connector parts onto the center conductor and braid. The contact is squeezed at the tip to secure the center conductor; but, if reuse is desired, the contact can be soldered. No braid soldering, braid combing, special crimping tools, or special adapters are needed. The result is a fail-safe, fast termination that eliminates faulty interconnections. Another advantage of the solderless termination is the absence of overheating during assembly and consequent damage to the cable itself. The Amphenol 83-58FCP offers performance equal to that provided by standard Amphenol 83-1SP-type uhf connectors, and at the same price.

At the heart of the solderless connecting mechanism is a body assembly featuring a hollow barrel with a barbed

end. After the cable is stripped, the slotted outer ferrule and coupling nut are slid onto the cable. Then the body assembly is pushed onto the cable dielectric but under the braid. The coupling nut is then slid over the body assembly, and the outer ferrule is pushed forward until it traps the cable braid against the rear flange of the body assembly. The ferrule then seats automatically. The resulting termination passed a 35-pound pull test.

For additional information about Amphenol 83-58FCP solderless connectors, contact Bunker Ramo RF Division, 33 East Franklin Street, Danbury, Connecticut 06810, or use *check-off* on page 142.

Bird Wattcher monitors rf power



Bird Electronic Corporation recently announced its new Series 3162 rf power monitor/alarm Wattcher,® designed to warn of rf power drop-off below a set level to conform with FCC part 21.107 requirements. This fast-acting (50 millisecond) unit can be used, for example, at a mobile terminal site to measure forward power, reflected power (with a momentary-contact front-panel switch), and to obtain VSWR and optimum system data during routine maintenance. It is then left in the line to feed back a signal in response to the transmitter being keyed, indicating whether it is on the air and with sufficient power.

The series 3162 is available for operation at all telephone company communications frequencies and power levels; for example, 2-512 MHz and 1-500 watts. It is available in either a 12-volt dc model or a 117-volt ac model. A typical 12-volt model is priced at $595.00.

For additional information, write Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139; telephone (216) 248-1200, or use *check-off* on page 142.

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